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## Chapter 22

### TESTING OF THREE-PHASE IMs

Experimental investigation or testing of induction machines at the manufacturer's and user's site may be considered an engineering art in itself.

It is also an indispensable tool in research and development of new induction machines in terms of new materials, sizing, topologies or power supply and application requirements.

There are national and international standards on the testing of IMs of low and large power with cage or wound rotor, fed from sinusoidal or PWM converter, and working in various environments.

We mention here the International Electrotechnical Committee (IEC) and the National Electrical Manufacturers Association (NEMA) with their standards on induction machines (IEC-34 series and NEMA MG1-1993 for large IMs).

Temperature, losses and efficiency, starting, unbalanced operation, overload, dielectric, cooling, noise, surge capabilities, and electromagnetic compatibility tests are all standardized.

A description of the standard tests is not considered here as the reader may study the standards for himself; the space required would be too large and the diversity of different standards prescriptions is so pronounced that it may create confusion for a newcomer in the field.

Instead, we decided to present the most widely accepted tests and a few non standardized ones which have recently been promoted with strong international vigor.

They refer to

- Loss segregation/power and temperature based methods
- Load testing/direct and indirect approaches
- Machine parameters estimation methods
- Noise testing methodologies

#### 22.1 LOSS SEGREGATION TESTS

Let us first recall here the loss breakdown ([Figure 22.1](#)) in the induction machine as presented in detail in Chapter 11.

For sinusoidal (power grid) supply, the time harmonics are neglected. Their additional losses in the stator and rotor windings and cores are considered zero.

However besides the fundamental, stator core and stator and rotor fundamental winding losses, additional losses occur. There are additional core losses (rotor and stator surface and tooth pulsation losses) due to slotting, slot-openings for different stator and rotor number of slots. Saturation adds new losses. Also, there are space harmonics produced time harmonic rotor-current losses which tend to be smaller in skewed rotors where surface iron additional losses are larger. The lack of insulation between the rotor cage-bars and the

rotor core allow for inter-bar currents and additional losses which are not negligible for high frequency rotor current harmonics.

All these nonfundamental losses are called either additional or stray load losses.

In fact, the correct term would be “additional” or “nonfundamental” as they exist, to some extent, even under no mechanical load. They accentuate with load and are, in general, considered proportional with current (or torque) squared.

On top of that, time-harmonics additional losses (Figure 22.1), both in copper and iron, occur with nonsinusoidal voltage (current) power supplies such as PWM power electronics converters.

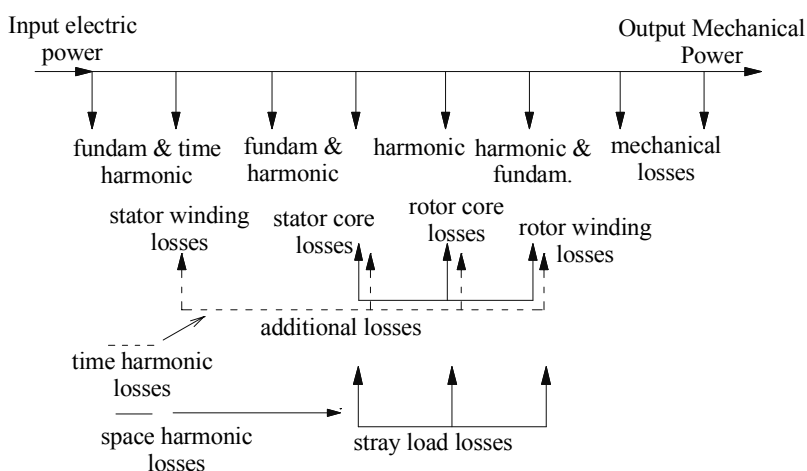


Figure 22.1 Loss breakdown in IMs

As a rather detailed analysis of such a complex loss composition has been done in Chapter 11, here we present only one sequence of testing for loss segregation believed to be coherent and practical. A few additional methods are merely suggested. This line of testing contains only the standard no load test at variable voltage, but extended well above rated voltage, and the stall rotor test.

The no load and shortcircuit (stall rotor) variable-voltage tests are known to allow for the segregation of mechanical losses,  $p_{mec}$ , the no load core losses and, respectively, the fundamental copper losses.

The extension of no load test well above rated voltage (as suggested in [1]) is used as a basis to derive an expression for the stray load losses considered proportional to current squared. The same tests are recommended for the PWM power electronics converter IM drives when the inverter is used in all tests.

### 22.1.1 The no-load test

A variable voltage transformer, with symmetric phase voltages, supplies an induction motor whose rotor is free at shaft. A data acquisition system acquires

three currents and three voltages and if available, a power analyzer, the power per each phase.

The method of two wattmeters leads to larger errors as the power factor on no load is low and the total power is obtained by subtracting two large numbers.

It is generally accepted that at no load, up to rated voltage, the loss composition is approximately

$$P_0 = 3R_1 I_{l0}^2 + p_{\text{iron}} + p_{\text{mec}} \quad (22.1)$$

If hysteresis losses are neglected or measured as the jump in input power when the motor on no load is driven through the synchronous speed, the iron losses may be assimilated with eddy current losses which are known to be proportional to flux and frequency squared. This is to say that  $p_{\text{iron}}$  is proportional to voltage squared:

$$p_{\text{iron}} = K_{\text{iron}} V_1^2 \quad (22.2)$$

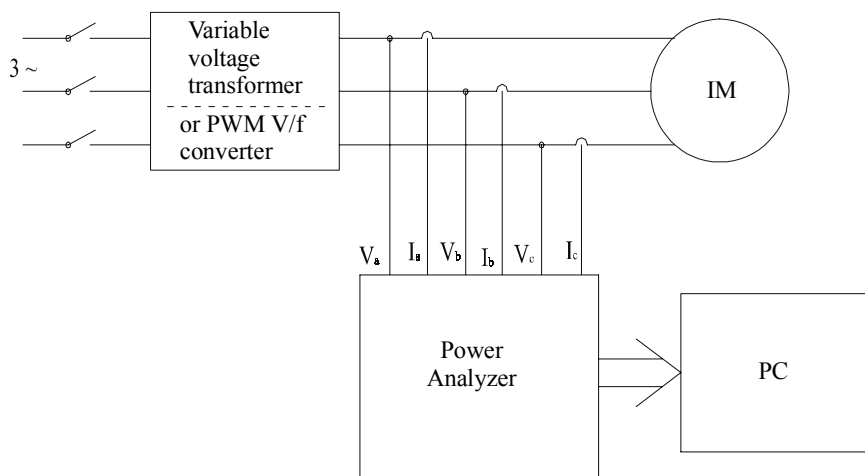


Figure 22.2 No-load test arrangement

When reducing the voltage to 25-30% of rated value, the speed decreases very little so the mechanical losses are independent of voltage  $V_1$ . The voltage reduction is stopped when the stator current starts rising.

Consequently,

$$P_0 - 3R_1 I_{l0}^2 = K_{\text{iron}} V_1^2 + p_{\text{mec}} \quad (22.3)$$

The stator resistance may be measured through a d.c. voltage test with two phases in series.

Alternatively, when all six terminals are available, the a.c. test with all phases in series is preferable as the airgap field is very low, so the core loss is negligible ([Figure 22.3](#)).

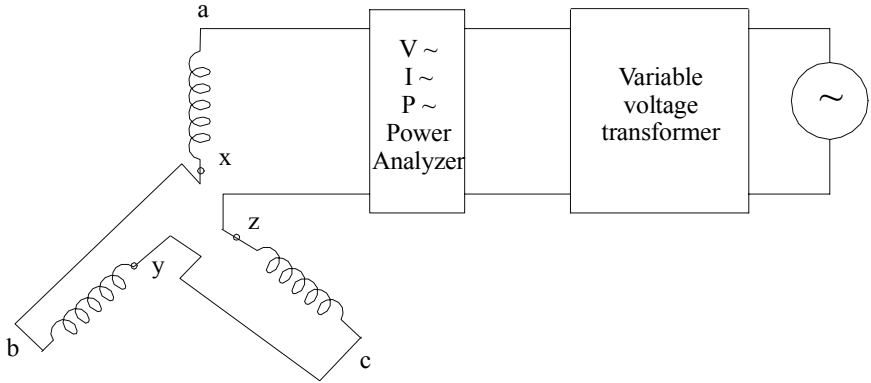


Figure 22.3 AC resistance measurement when six terminals are available

With voltage, current and power measured, the stator resistance  $R_s$  and homopolar reactance  $X_0'$  (lower or equal to stator leakage reactance) are:

$$R_{1\sim} = \frac{V_{\sim}}{3I_{\sim}} \quad (22.4)$$

$$X_0 \leq X_{1l} = \sqrt{\left(\frac{V_{\sim}}{3I_{\sim}}\right)^2 - R_1^2} \quad (22.5)$$

With full pitch coil windings,  $X_0 = X_{1l}$ . However,  $X_0 < X_{1l}$  for chorded windings. Low voltage is required to avoid over-currents in this test.

For large machines with skin effect in the stator, even at fundamental frequency,  $R_{1\sim}$  is required.

The same is valid with IMs fed from PWM power converters. In the latter case, the Variac is replaced by the PWM converter, triggered for two power switches only, with a low modulation index.

The graphical representation of (22.3) is shown on [Figure 22.4](#).

The intercept of the graph on the vertical axis is the mechanical losses.

The fundamental iron losses  $p_{iron}^1$  result from the graph in [Figure 22.4](#) for various values of voltage.

We may infer that in most IMs the stator impedance voltage drop is small so the fundamental core loss determined at no load is valid as well for on load conditions.

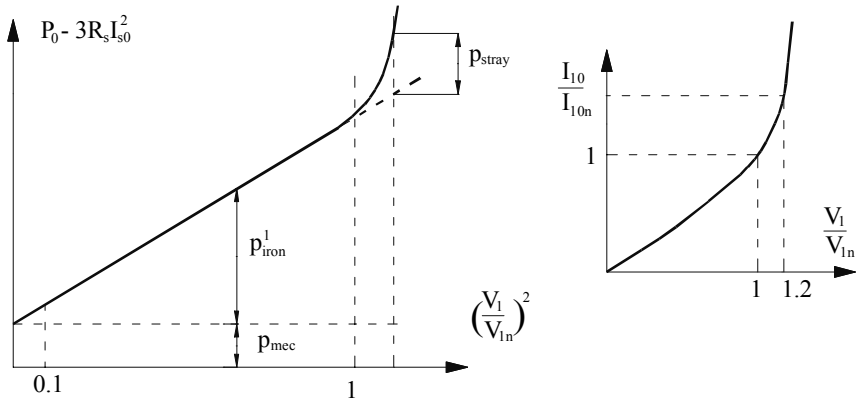


Figure 22.4 No load input (less stator copper loss) versus voltage squared

In reality under load conditions, the value of  $p_{iron}^1$  slightly decreases as the e.m.f. does the same.

### 22.1.2 Stray losses from no-load overvoltage test

When increasing the voltage over its rated (design) value, Equation (22.3) depart from a straight line (Figure 22.4), at least for small power induction machines. [1]

Apparently in this case, the iron saturates so the third flux harmonic due to saturation causes more losses. This is not so in most IMs as explained in paragraph 5.4.4. However, the stator current increases notably above rated no load current. Space harmonics induced currents in the rotor will increase also. Not so for the fundamental rotor current. So the stator core surface and tooth pulsation losses are not notable. Fortunately, they are not large even under load conditions.

All in all, it is very tempting to use this extension of no load test to find the stray loss coefficient as

$$K_{stray} = \frac{P_{stray}}{3I_{10}^2} \quad (22.6)$$

With the voltage values less than 115-120%, the difference in power on Figure 22.4 is calculated, together with the respective current  $I_{10}$  measured. A few readings are done and an average value for the stray load coefficient  $K_{stray}$  is obtained.

Test results on three low power IMs (250 W, 550 W, 4.2 kW) have resulted in stray losses at rated current of 2%, 0.9%, and 2.5%. [1] It is recognized that stray load losses, generalizations have to be made with extreme caution.

On load tests are required to verify the practicality of this rather simple method for wide power ranges. It is today recognized that stray load losses are much larger than 0.5% or 1% as stipulated in some national and in IEC

standards. [3] Their computation from on load tests, as in IEEE 112B standard, seems a more realistic approach. As the on load testing is rather costly, other simpler methods are considered.

Historically, the reverse rotation test at low voltage and rated current has been considered an acceptable way of segregating stray load losses. [2]

### 22.1.3 Stray load losses from the reverse rotation test

Basically, the IM is rotated in the opposite direction of its stator travelling field (Figure 22.5). The stator is fed at low voltage through a Variac.

With the speed  $n = -f_1/p$ , the value of slip  $S = 1 - np_1/f_1 = 2$ , and thus the frequency of rotor currents is  $2f_1$ .

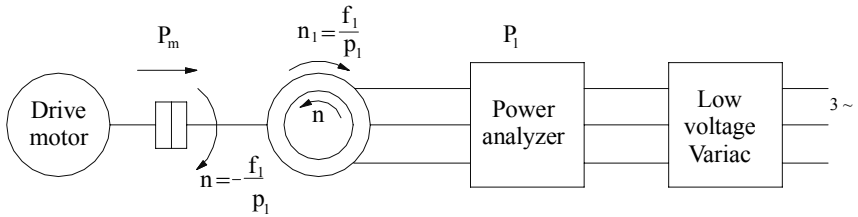


Figure 22.5 Reverse rotation test

The mechanical input will cover the mechanical losses  $p_{mec}$ , the rotor stray losses plus the term due to  $3I_2^2R_2(1 - S)/S$ . For  $S = 2$  this term is  $-3I_2^2R_2/2$ . Consequently, the difference between the stator electric input  $P_1$  and the mechanical input  $P_m$  is

$$P_1 - P_m = 3R_1I_1^2 + P_{iron} + 3\frac{I_2^2R_2}{2} - P_{stray} - \frac{3I_2^2R_2}{2} - P_{mec} \quad (22.7)$$

The rotor winding loss terms cancel in (22.7).

The fundamental iron losses in  $P_{iron}$  are different from those for rated power motoring as they tend to be proportional to voltage squared. Also  $P_{iron}$  contains the stator flux pulsation losses, which again may be considered to depend on voltage squared.

So,

$$P_{iron} \approx P_{(iron)rated} \cdot \left( \frac{V_1}{V_{1n}} \right)^2 \quad (22.8)$$

The method has additional precision problems as detailed in [4] besides the need for a drive with measurable shaft torque (power).

By comparison, the extension of no load test above rated voltage is much more practical. But is it a satisfactory method? Only time will tell as many other attempts to segregate the stray load losses have not yet gotten universal acceptance.

### 22.1.4 The stall rotor test

Traditionally, the stall rotor (shortcircuit) test is done with a three phase supply and mechanical blockage to stall the rotor. With single phase supply, however, the torque is zero and thus the rotor remains at standstill by itself (Figure 22.6).

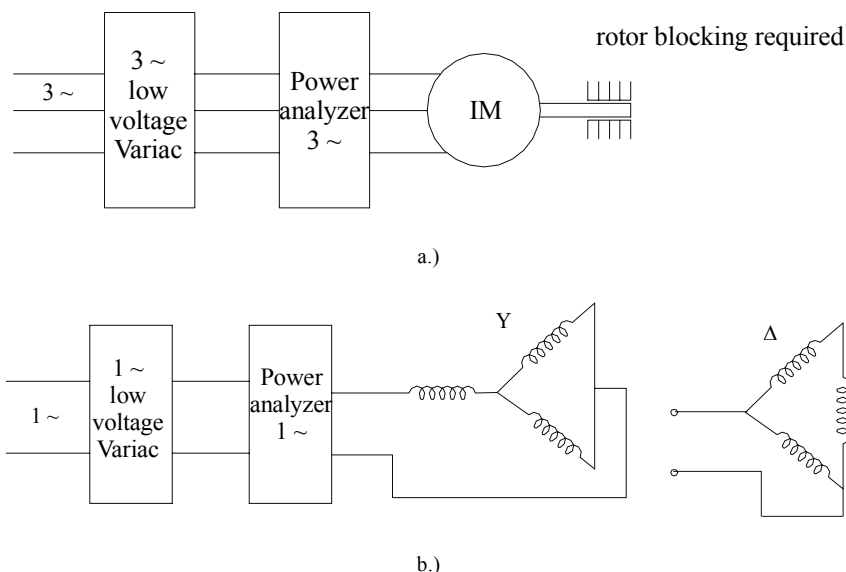


Figure 22.6 Stall rotor tests

a.) three phase supply

b.) single phase supply

The tests are done for low voltages until the current reaches its rated value. If the test is done at rated frequency  $f_{ln}$ , the skin effect in the rotor is pronounced and thus the rotor resistance is notably larger than in load operation when the slip frequency  $Sf_{ln} \ll f_{ln}$ . Only for low power IMs (in the kW range), of general design (moderate starting torque), the assumption of low rotor skin effect is true.

It is true also for all wound rotor IMs.

With voltage, current, and power measured, the stall rotor (shortcircuit) resistance  $R_{sc}$  and reactance  $X_{sc}$  are found.

We use the superscript  $s$  to emphasize that these values have been obtained at stall. The values of  $R_{sc}^s$  and  $X_{sc}^s$  are less practical than they seem for load conditions. Skin effect and leakage saturation, at high values of stall currents, for rated voltage, lead to different values of  $R_{sc}^s$ ,  $X_{sc}^s$  at stall.



$$\begin{aligned}
R_1 + R_2^s &= R_{sc}^s = \frac{P_{sc3}}{3I_{sc3}^2}; \\
X_{sc}^s &= \sqrt{\left(\frac{V_{sc3}}{I_{sc3}}\right)^2 - R_{sc}^2} = X_{11}^s + X_{12}^s \\
R_1^s + R_2^s &= R_{sc}^s = 2 \frac{P_{sc1}}{3I_{sc1}^2}; \\
X_{sc}^s &= \sqrt{\frac{2}{3} \left(\frac{V_{sc1}}{I_{sc1}}\right)^2 - R_{sc}^2} = X_{11}^s + X_{12}^s
\end{aligned} \tag{22.9}$$

The shortcircuit test is also supposed to provide for fundamental winding losses computation at rated current,

$$P_{Co} = P_{co1n} + P_{co2n} = 3R_{sc}^s \cdot I_{1n}^2 \tag{22.10}$$

That is, to segregate the winding losses for rated currents. It manages to do so correctly only for IMs with no skin effect at rated frequency in the rotor ( $Sf_1 = f_1$ ) and in wound rotor IMs.

However, if a low frequency low voltage power source is available (even 5% of rated frequency will do), the test will provide correct data of rated copper losses (and  $R_1+R_2$ ,  $X_{11}+X_{12}$ ) to be used in fundamental winding losses for efficiency calculation attempts.

### 22.1.5 No-load and stall rotor tests with PWM converter supply

The availability of PWM converters for IM drives raises the question if their use in no – load and stall tests as variable voltage and frequency power sources is not the way of the future.

It is evident that for IMs destined for variable speed applications, with PWM converter supplies, the no load and stall tests are to be performed with PWM converter rather than Variac (transformer) supplies.

As the voltage supplied to the motor has time harmonics, the stator and rotor currents have time harmonics. So, in the first place, even if  $S \approx 0$ , the time harmonics produced rotor currents are non zero as their slip is  $S_v \approx 1$ . The space harmonics produced rotor harmonics currents, existing also with sinusoidal voltage supply, are augmented by the presence of stator voltage time harmonics. They are, however, load independent ( $S_v \approx 1$ ).

So the loss breakdown with PWM converter supply at no load is

$$P_o' = 3R_1' I_{10}^{\prime 2} + 3R_2' I_{20}^{\prime 2} + p_{iron}' + p_{mec} \tag{22.11}$$

It has been shown that, despite the fact that the stator and rotor no load currents show time harmonics due to PWM converters, the airgap e.m.f. is quasi-sinusoidal. [5] And so is the magnetization current  $I_m$ .

So, performing the no load test for the same fundamental voltage and frequency, once with sinusoidal power source and once with PWM converter, the current relationships are

$$\begin{aligned} I_{10}^2 &= I_m^2 \\ I_{10}^2 &\approx I_{20}^2 + I_m^2 \end{aligned} \quad (22.12)$$

Thus the rotor current under no load  $I'_{20}$  for PWM converter supply is

$$I'_{20} = \sqrt{I_{10}^2 - I_m^2} \quad (22.12')$$

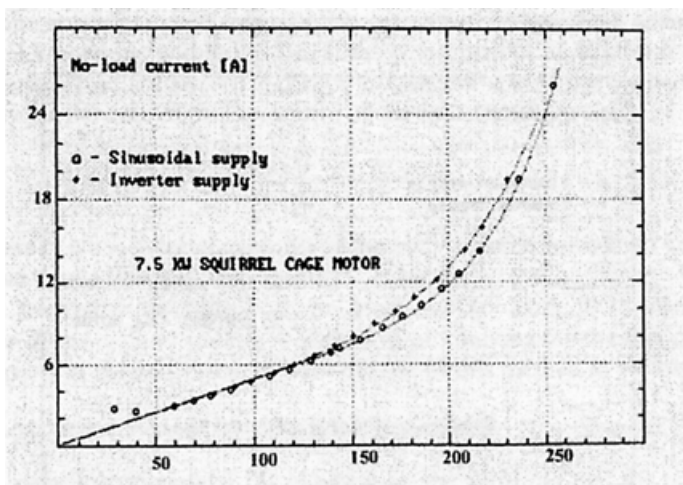
The stator and rotor resistances  $R'_1$  and  $R'_2$  valid for the given current harmonics spectrum are not known.

However, with the rotor absent, a slightly overestimated value of  $R'_1$  may be obtained. From a stall rotor test at low 5% fundamental frequency,  $R'_1 + R'_2$  may be found.

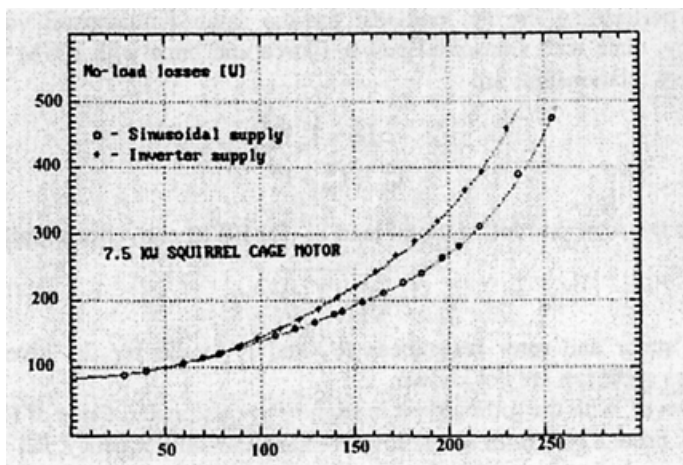
From Equation (22.11), we may represent graphically  $p'_{iron} + p'_{mec}$  as a function of fundamental voltage squared for various fundamental frequencies. Finally, we obtain  $P_{iron}(f_1, (V_1/V_{1n})^2)$  and  $p_{mec}(f_1)$ .

In general, for PWM converter supply, both the no load losses and the stall rotor losses are larger than for sinusoidal supply with same fundamental voltage and frequency.

With high switching frequency PWM converters, the time harmonics content of voltage and current is less important and thus smaller no load and stall rotor additional losses occur [5] (Figure 22.7-22.8). For the cage rotor IM stall tests, the difference in losses is negligible, though the current (RMS) is larger (Figure 22.8)



a.)



b.)

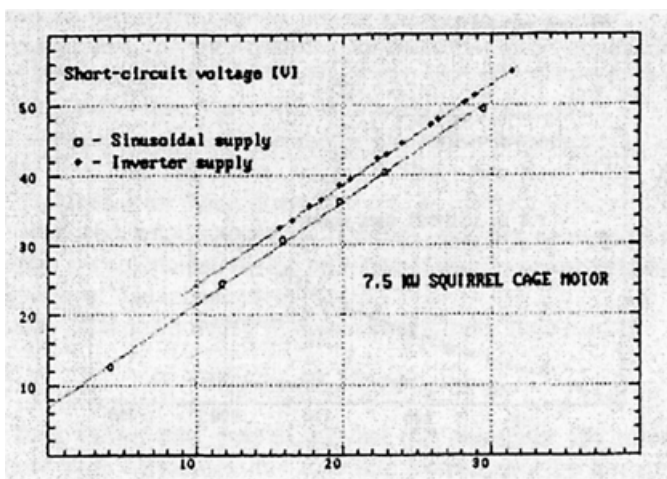
Figure 22.7 No load testing

a.) current; b.) losses

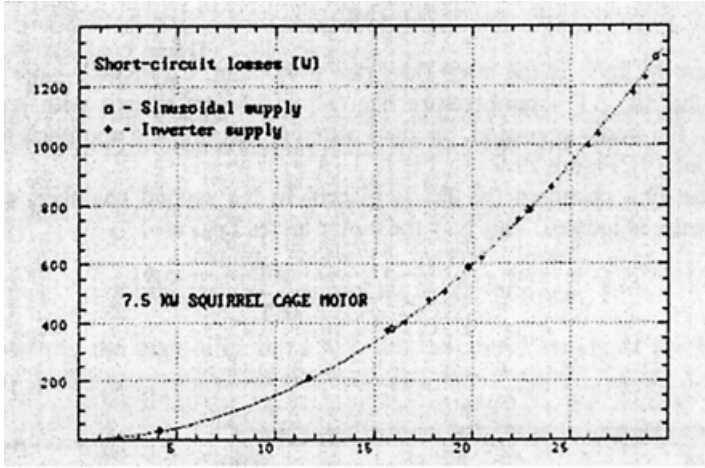
It may be argued that for given a.c. power grid parameters, in general, PWM converters cannot produce rated voltage fundamental to the motor.

A 3-5% voltage fundamental reduction due to converter limits is accepted. So either correction of all losses, proportional to voltage ratio squared, is applied to PWM converter tests or the power grid is set to provide 5% more than rated voltage fundamental.

As for the extension of the voltage beyond rated voltage by 10-15% for the no load test, to calculate the stray load loss coefficient mentioned in a previous paragraph, a voltage up/down transformer is required. Alternatively, the ratio  $V_{in}/f_1$  may be increased by decreasing the frequency.



a.)



b.)

Figure 22.8 Stall rotor test  
a.) current b.) losses

All this being said, it appears that the no load and stall rotor tests for loss segregation may be extended to PWM converter supplied induction motors.

Pertinent data acquisition and processing systems are required with non-sinusoidal voltages and currents to calculate fundamentals and losses.

In parallel with the electrical methods to segregate the losses in IM, presented so far, temperature/time and calorimetric methods have also been developed to determine the various losses in the IM.

Temperature-time methods require numerous temperature sensors to be planted in key points within the IM. If the loss distribution is known and IM is disconnected from the source and kept at constant speed, the temperature/time derivative, at the time of disappearance of the loss source, is

$$\frac{dT}{dt} = -\frac{1}{C} \cdot Q \quad (22.13)$$

Where C is the thermal capacity of the body volume considered and Q – the respective whole losses in that region.

The tests may be done both on load and no load. [6,7] The intrusive character of the method and the requirement of knowing the thermal capacity of various parts of IM body seem to limit the use of this method to prototyping.

#### 22.1.6 Loss measurement by calorimetric methods

The calorimetric method [8] is based on the principle that the temperature rise in a wall insulated chamber is proportional to the losses dissipated inside.

For steady state (thermal equilibrium), the rate of heat transfer by the air coolant, Q, which in fact represents the IM losses, is

$$Q = MC_p \Delta T \quad (22.14)$$

Where  $M(\text{kg/s})$  is the mass flow rate of coolant,  $C_p(\text{J/KgK})$  – the specific heat of the air;  $\Delta T$  – temperature rise (K) of coolant at exit with respect to entrance. For better precision, the dual chamber calorimetric approach has been introduced. [9], [Figure 22.9](#)

In the first chamber, the IM is placed. In the second chamber, a known power heater is located. This way the motor losses  $\Sigma p_{\text{IM}}$  are

$$\Sigma p_{\text{IM}} = P_{\text{heater}} \times \frac{\Delta T_1}{\Delta T_2} \quad (22.15)$$

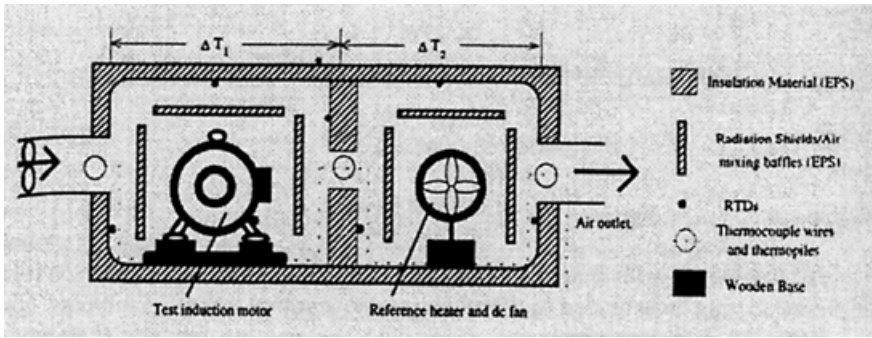


Figure 22.9 The dual chamber calorimeter method

The air properties, not easy to find with variable temperature, are not required. Also the errors of temperature sensors tend to cancel each other, as we need only  $\Delta T_1 / \Delta T_2$ .

There is some heat leakage from the calorimeter. This loss may be determined with a precision of  $\pm 0.5\text{W}$ . [9] Also, as an order of magnitude, 60-100 litres/sec of air is required for loss measurements of 0.2-1 kW. A proper temperature rise per chamber should be around 10C.

The overall uncertainty of the method is about  $\pm 15\text{ W}$ . The friction losses due to stuffing the motor shaft through the chamber walls have to be considered.

For IMs with losses above 150 W, an error in the loss measurements less than 10% is expected. The method works for any load levels and with any kind of IM supply. However, it takes time – until thermal steady state is reached and implies notable costs. For prototyping, however, it seems appropriate.

## 22.2 EFFICIENCY MEASUREMENTS

Due to its impact on energy costs, efficiency is the single most important parameter index in electric machines.

As induction motors are fabricated and used worldwide, national and international standards for efficiency measurements have been introduced. We mention here some of them which are deemed to be highly representative

- IEEE Standard 112–1996 [10]
- IEC–34–2 and IEC–34–2A [11]
- JEC 37

NEMA MG1–1993 and the Canadian standard C390 correspond to IEEE–112 while most European countries abide by IEC standard. JEC holds in Japan, mainly.

The definition of efficiency  $\eta$  is basically unique

$$\eta = \frac{\text{output\_power}}{\text{input\_power}} = 1 - \frac{\text{overall\_losses}}{\text{input\_power}} \quad (22.16)$$

However, the main difference between the above standards involves how the stray load losses are defined and treated as part of overall losses.

Direct measurement of stray losses is obtained from

$$P_{\text{stray}} = P_{\text{input}} - (P_{\text{output}} + p_{\text{iron}} + 3R_1 I_1^2 + 3R_2 I_2^2 + p_{\text{mec}}) \quad (22.17)$$

$$\text{with: } I_2 \approx \sqrt{I_1^2 - I_m^2}$$

where:  $I_m$ , the magnetization current, is equal to the no load current at the respective voltage.

The various losses in (22.17) are obtained through the loss segregation methods. The input and output powers are measured directly.

Let us now summarize how the efficiency and stray load losses are handled in the IEEE–112 and IEC–34–2 standards.

### 22.2.1 IEEE Standard 112–1996

This rather complete standard consists of five methods to determine efficiency. They are called methods A, B, C, E and F.

Method A is, in fact, a direct method where the input and output powers are measured directly.

Method B directly measures the stray load losses.

Linear regression is used to reduce measurement errors. The stray load losses are considered proportional to torque squared.

$$(P_{\text{stray}})_{\text{corrected}} = AT_{\text{shaft}}^2 \quad (22.18)$$

The correlation coefficient of linear regression has to be larger than 0.9 to secure good measurements.

Method C introduces a back to back (motor/generator) test. The stray load losses are obtained by loss segregation (22.16) but the measured power is, in fact, the difference between the input and output of the two identical machines. So the total losses in the two machines are measured. Consequently, better precision is expected. The total stray load losses thus segregated are divided between the motor and generator considering that they are proportional to current squared.

Method E and  $E_1$  are indirect methods. So the output power is not measured directly.

**Table 22.1 Assumed stray load losses/rated output/IEEE–112 method E1**

| Rated power (kW) | Stray load losses |
|------------------|-------------------|
| 0.75 – 90        | 1.8%              |
| 91 – 375         | 1.5%              |
| 376 – 1800       | 1.2%              |
| $\geq 1800$      | 0.9%              |

In fact, in method E, the stray load losses are measured via the reverse rotation test (see paragraph 22.1.3). In method  $E_1$ , the stray load losses are simply assumed at a certain value (Table 22.1).

Methods F and  $F_1$  make use of the equivalent circuit with stray losses directly measured (F) or assigned a certain value ( $F_1$ ).

### 22.2.2 IEC standard 34–2

The efficiency is estimated by determining all loss components of the IM.

The losses are to be determined by the loss segregation methods or from the measurement of overall losses

- Direct load test with torque measurement
- Calibrated torque load machine
- Mechanical back to back test as for IEEE–112 C
- Electrical back to back test

The preferred method in the standard is however the segregation of losses with stray load losses having a fix value of 0.5% of rated power. The Japanese standard JEC neglects the stray load losses altogether.

Stray load losses in Table 22.1 (IEEE–112E1) are notably higher than the 0.5% in IEC–34–2 standard and the zero value in JEC standard. Consequently, the same induction motor would be labelled with highest efficiency in the JEC standard, then in IEC–34–2 and finally the lowest, and most realistic in the IEEE–112E1 standard.

### 22.2.3 Efficiency test comparisons

Typical tests according to the three standards run on the same induction motor of 75 kW are shown in Table 22.2. [12]

A few remarks are in order

- The first test, the direct method, involves direct measurement of input electrical power, shaft torque and speed with a total possible error in efficiency of maximum 1%.
- IEEE–172B tests imply that linear regression is used on the measured stray losses, considered proportional to torque squared.
- For IEC–34–2 the stray load losses are considered 0.5% of rated output and proportional to torque squared.

**Table 22.2 Efficiency testing of a 75 kW standard IM according to direct measurements: IEEE-112B, IEC-34-2 and JEC**

|                                                                                                            |       |       |       |       |
|------------------------------------------------------------------------------------------------------------|-------|-------|-------|-------|
| Load                                                                                                       | 25%   | 50%   | 75%   | 100%  |
| $U_{\text{average}}$ [V]                                                                                   | 407.8 | 404.7 | 400.2 | 391.3 |
| $I_{\text{average}}$ [A]                                                                                   | 71.4  | 91.2  | 118.1 | 149.6 |
| $P_{\text{el}}$ [W]                                                                                        | 22270 | 41820 | 61850 | 81360 |
| Power Factor [-]                                                                                           | 0.44  | 0.65  | 0.76  | 0.80  |
| $T$ [Nm]                                                                                                   | 182.2 | 364.0 | 548.3 | 723.3 |
| $n$ [rpm]                                                                                                  | 997   | 995   | 994   | 991   |
| $P_{\text{shaft}}$ [W]                                                                                     | 19021 | 37924 | 57078 | 75064 |
| slip [%]                                                                                                   | 0.3   | 0.5   | 0.6   | 0.9   |
| $P_{\text{loss}}$ measured directly [W]                                                                    | 3249  | 3896  | 4772  | 6296  |
| $P_{R1.1}$ [W]                                                                                             | 354   | 578   | 968   | 1553  |
| $P_{\text{core}}$ [W]                                                                                      | 1774  | 1728  | 1660  | 1527  |
| $P_{R1.2}$ [W]                                                                                             | 60.4  | 198   | 355   | 705   |
| $P_{\text{w.fr}}$ [W]                                                                                      | 910   | 910   | 910   | 910   |
| $P_{\text{stray}}$ [W]                                                                                     | 151   | 482   | 879   | 1601  |
| $P_{\text{shaft}}/P_{\text{rated}}$ [%]                                                                    | 25.4  | 50.6  | 76.1  | 100.1 |
| <b>Efficiency Direct [%]</b>                                                                               | 85.4  | 90.7  | 92.3  | 92.3  |
| Efficiency using IEEE-112<br>Linear regression coefficient. Slope: $2.802 \cdot 10^{-3}$ Correlation: 0.99 |       |       |       |       |
| $P_{\text{stray corr.}}$ [W]                                                                               | 93    | 371   | 843   | 1466  |
| $P_{\text{loss core}}$ [W]                                                                                 | 3191  | 3785  | 4736  | 6161  |
| $P_{\text{shaft core}}$ [W]                                                                                | 19079 | 38035 | 57114 | 75199 |
| $P_{\text{shaft}}/P_{\text{rated}}$ [%]                                                                    | 25.4  | 50.7  | 76.2  | 100.3 |
| <b>Efficiency IEEE – 112 [%]</b>                                                                           | 85.7  | 90.9  | 92.3  | 92.4  |
| Efficiency using IEC 34-2                                                                                  |       |       |       |       |
| $P_{\text{stray}}$ [W]                                                                                     | 93    | 151   | 253   | 407   |
| $P_{\text{loss}}$ [W]                                                                                      | 31.91 | 3565  | 4146  | 5102  |
| $P_{\text{shaft}}$ [W]                                                                                     | 19079 | 38255 | 57704 | 76258 |
| $P_{\text{shaft}}/P_{\text{rated}}$ [%]                                                                    | 25.4  | 51.0  | 76.9  | 101.7 |
| <b>Efficiency IEC 34 – 2 [%]</b>                                                                           | 85.7  | 91.5  | 93.3  | 93.7  |
| Efficiency using JEC                                                                                       |       |       |       |       |
| $P_{\text{stray}}$ [W]                                                                                     | 0     | 0     | 0     | 0     |
| $P_{\text{loss}}$ [W]                                                                                      | 3098  | 3414  | 3893  | 4695  |
| $P_{\text{shaft}}$ [W]                                                                                     | 19172 | 38406 | 57957 | 76665 |
| $P_{\text{shaft}}/P_{\text{rated}}$ [%]                                                                    | 25.6  | 51.2  | 77.3  | 102.2 |
| <b>Efficiency JEC [%]</b>                                                                                  | 86.1  | 91.8  | 93.7  | 94.2  |

The main conclusion is that differences in efficiency, with respect to direct measurements of more than 2% may be encountered with the IEC–34–2 and JEC, while IEEE–112B is much closer to reality.



The evident suggestion then is to use the IEEE–112B method as often as possible whenever the direct method is feasible. The problem is that the direct method requires almost one day testing time per motor, notable man power and energy.

Providing and mounting a brushless torque-meter with integrated speed sensor both of high precision 1% and, respectively, 1 rpm speed error, is not an easy task as torque (power) increases. Also the power grid rating increases with the power of the tested IM.

### 22.2.4 The motor/generator slip efficiency method

Difficulties related to the torque-meter acquisition mounting and frequent calibration and excessive energy consumption may be tamed by eliminating the torque-meter and loading the IM with another IM fed from the now available bi-directional PWM voltage-source converter supply used for high performance variable speed drives (Figure 22.10). This method may be seen as an extension of IEEE–112C back to back method. A very precise speed sensor or a slip frequency measuring device is still required.

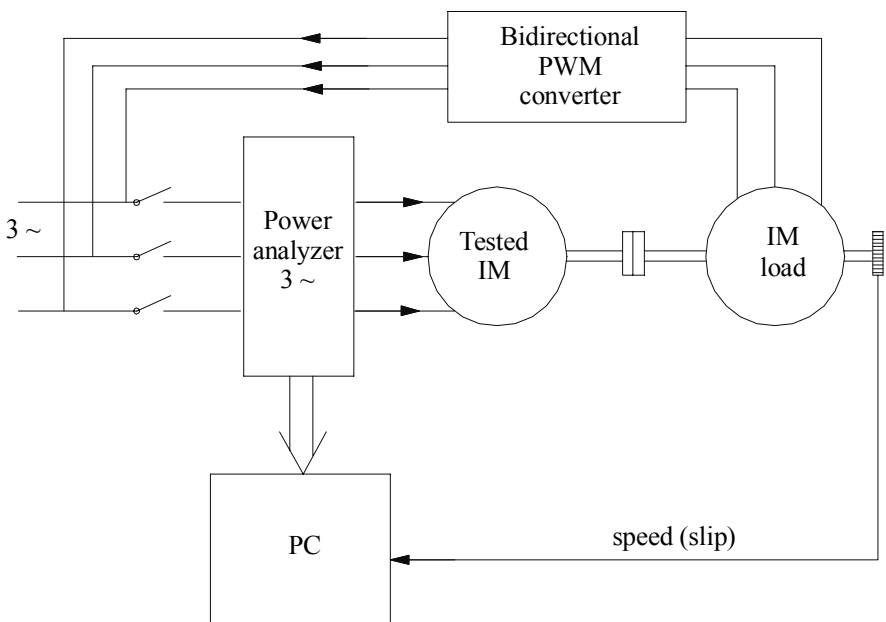


Figure 22.10 Motor/generator slip efficiency method [13]

In principle, the IM is tested for  $\pm 0.25$ ,  $\pm 0.5$ ,  $\pm 0.75$ ,  $\pm 1.0$ ,  $\pm 1.25$ ,  $\pm 1.5$  times rated slip speeds as a motor and as a generator, respectively, with the voltages, currents, and input power measured via a power analyzer.

It is admitted that the power level is proportional to slip so the 25%, 50%, 75%, 100%, 125%, 150% loads are approximately obtained. Alternatively, the slip is adjusted to the required power. The power balance equations for the motor and generator modes and equal (or known) slip values are

$$P_{in/m} = 3R_1 I_{1m}^2 + p_{iron/m} + K_{stray} \cdot I_{1m}^2 + 3R_2 \frac{I_{2m}^2}{S_m} \quad (22.19)$$

$$P_{out/g} = 3 \frac{R_2 I_{2g}^2}{|S_g|} - K_{stray} I_{1g}^2 - p_{iron/g} - 3R_1 I_{1g}^2 \quad (22.20)$$

For the 25%-150% load range the rotor circuit is highly resistive and thus,

$$I_{2m}^2 = I_{1m}^2 - I_m^2 \quad (22.21)$$

$$I_{2g}^2 = I_{1g}^2 - I_m^2 \quad (22.22)$$

With  $I_m$  equal to the no load current at rated voltage, the rotor currents for motor and generator modes,  $I_{2m}$  and  $I_{2g}$  are obtained.

The core losses are considered with their value at no load and rated voltage, the same for motor and generator,

$$P_{iron/m} = P_{iron/g} \quad (22.23)$$

Now in Equations (22.19)–(22.20) we have only two unknowns:  $R_2$ –rotor resistance and  $K_{stray}$  (stray load coefficient). As the equations are linear, the solution is straightforward.

It may be argued that while  $R_2$  enters large power components,  $K_{stray}$  enters loss(small) power components. While this is true, the direct method again subtracts output from input to compute efficiency. Good instrumentation should deal with this properly.

A few remarks on this method now follow:

- No torque-meter is required
- Motor/generator slip loading is performed
- The motor/generator slip levels may be rather equal but accuracy is not a must
- The core losses and the no load current at rated voltage is required
- The mechanical losses  $p_{mec}$ , as determined from variable voltage no load test, are required to calculate the efficiency

$$\eta = 1 - \left( 3R_1 I_{1m}^2 + p_{iron/m} + 3K_{stray} I_{1m}^2 + 3R_2 I_{2m}^2 + p_{mec} \right) / P_{in/m} \quad (22.24)$$

Efficiency could also be calculated from generator mode test results with

$$p_{mecg} = p_{mecm} \cdot (1 + |S|)^2 \quad (22.25)$$

The rotor resistance  $R_2$  at low (slip) frequency may be determined eventually only at 25% rated slip and considered constant over the loading tests at 50%, 75%, 100%, 125%, 150% as the temperature also has to stabilize for each test. This way for all loads except for 25% only the difference between Equations (22.19)–(22.20) will be used. This should enhance precision notably. This estimation of  $R_2$ , independent of skin effect, should be preferred to that from stall rotor tests at rated frequency. To cut the testing time, we may acquire the temperature of the stator frame at the time of every load level testing and correct the stator and rotor resistances accordingly. The stator resistance is to be found from a d.c. test at a known temperature.

If time permits, the machine may be stopped quickly, by fast braking it with the load machine which is supplied by a bi-directional power flow converter, for the stator resistance to be d.c. (or miliohmeter) measured.

Test results in [13] proved satisfactory when compared with IEEE–112B tests.

Still the coupling of a loading machine at the IM shaft is required. For vertical shaft IMs as well as for large power IMs, this may be incurring too high costs.

Recently, the mixed frequency method, stemming from the classical two-frequency method [14], has been revived by using a static power converter to supply the IM. [15,16] The key problem in this artificial loading test is the equivalence of losses with direct load testing.

### 22.2.5 The PWM mixed frequency temperature rise and efficiency tests

The two-frequency test relies on the principle of supplying the IM with voltages  $V_1$ ,  $V_2$  of two different frequencies  $f_1$  and  $f_2$

$$V(t) = V_1 \cos \omega_1 t + V_2 \cos \omega_2 t \quad (22.26)$$

The rated frequency  $f_1$  differs from  $f_2$  such that the IM speed oscillates very little. Consequently, the IM works as a motor for  $f_1$  and as a generator for  $f_2$ . The RMS value of stator current depends on the amplitude and frequency of second voltage:  $V_2$ ,  $f_2$ .

Traditionally, the test uses a synchronous generator of frequency  $f_2 \approx (0.8–0.85)f_1$  and  $V_2 < V_1$  connected in series with the power grid voltage  $V_1(f_1)$  to supply the IM.  $V_2$ ,  $f_2$  are modified until the RMS value of stator current has the desired value (around or equal to rated value in general).

The test has been traditionally used to determine the temperature rise transients and steady state value, to replicate direct load loss conditions. The exceptional advantage is that no loading machine is attached to the shaft. Consequently, the testing costs and time are drastically reduced, especially for vertical shaft IMs.

However, due to the lack of correct loss equivalence such tests were reported to yield, in general 6–10°C higher temperature than direct load tests for same RMS current.

This is one reason why the mixed frequency traditional tests did not get enough acceptance.

The availability of PWM converters should change the picture to the point that such tests could be applied, at least for IMs destined for variable speed drives, to determine the efficiency. There are quite a few ways to control the PWM converter to obtain artificial loading.

Among them we mention here,

- The accelerating–decelerating method [15]
- Fast primary frequency oscillation method [15]
- PWM dual frequency method [16]
- The accelerating–decelerating method

The IM is supplied from an off the shelf PWM converter – even one with unidirectional power flow will do. The reference speed is ramped up and down linearly or sinusoidally around the rated frequency synchronous speed so that the IM with free shaft can follow the path.

The speed variation range  $n_{\min} - n_{\max}$  and its time variation:

$$\frac{dn}{dt} \approx \frac{n_{\max} - n_{\min}}{T} \quad (22.27)$$

are used as parameters to try to provide loss equivalence with the direct load test.

Common sense indicates that for equivalent stator copper loss, the same RMS current as in direct load testing is required.

As the speed has to go above rated (base) speed, the inverter voltage ceiling is causing constant voltage to be applied above rated speed with only frequency as a variable.

As  $(n_{\max} - n_{\min})/n_{\text{rated}} \approx 0.2 - 0.4$ , the mechanical losses vary. As they are, in general, proportional to speed squared, their equivalencing should not be very difficult.

$$(P_{\text{mec}})_{\text{rated}} = K_m n_n^2 = \frac{1}{(n_{\max} - n_{\min})} \int_{n_{\min}}^{n_{\max}} K_m n^2 dn = \frac{n_{\max}^3 - n_{\min}^3}{3(n_{\max} - n_{\min})} \cdot K_m \quad (22.28)$$

Consequently,

$$n_{\max}^2 + n_{\max} \cdot n_{\min} + n_{\min}^2 = 3n_n^2$$

As the frequency varies, all core losses are considered proportional to voltage squared, and the torque is also proportional with voltage squared (constant slip during the tests), the average core loss is slightly smaller than for direct load tests. In general,  $n_{\max}$  and  $n_{\min}$  are rather symmetric with respect to rated speed.

Consequently,  $n_{\max}$  and  $n_{\min}$  are found easily.

Now the pace of acceleration is left to be adjusted so as to produce the equivalent RMS stator current over the speed oscillation cycle.

An RMS current estimator based on current acquisition over a few periods can be used as current feedback signal. The reference RMS current will be input to a slow current close loop that outputs the reference speed oscillation frequency between  $n_{\min}$  and  $n_{\max}$ .

The voltages  $V_a$ ,  $V_b$  and the currents  $I_a$  and  $I_b$  are acquired and transformed to synchronous coordinates

$$\begin{aligned} \begin{pmatrix} V_d(I_d) \\ V_q(I_q) \end{pmatrix} &= \frac{2}{3} \begin{pmatrix} \cos(-\theta_{es}) & \cos\left(-\theta_{es} + \frac{2\pi}{3}\right) & \cos\left(-\theta_{es} - \frac{2\pi}{3}\right) \\ \sin(-\theta_{es}) & \left(-\theta_{es} + \frac{2\pi}{3}\right) & \sin\left(-\theta_{es} - \frac{2\pi}{3}\right) \end{pmatrix} \cdot \begin{pmatrix} V_a(I_a) \\ V_b(I_b) \\ V_c(I_c) \end{pmatrix}; \\ V_c &= -V_a - V_b; \\ I_c &= -I_a - I_b; \\ \theta_{es} &= \int \omega_1 dt; \end{aligned} \quad (22.29)$$

The frequency  $f_1$  varies with time as the machine accelerates and decelerates.

The input power

$$P_1 = \frac{3}{2} (V_d I_d + V_q I_q) \quad (22.30)$$

is in fact the instantaneous active power.

If the reference frame axis  $d$  falls along phase  $a$  axis at time zero, then  $V_q = 0$  and thus only one term appears in (22.30).

The reactive input power  $Q_1$  is

$$Q_1 = \frac{3}{2} (V_d I_q - V_q I_d) \quad (22.30')$$

and retains also one term only. Notice that the current has both components.

The main advantage of this variable transformation – to be done off line through a PC – is that it provides “instantaneous” power values without requiring averaging over a few integral electrical periods. [15]

Consequently, while the machine travels from  $n_{\min}$  to  $n_{\max}$  and back, the active and (reactive) instantaneous power travel a circle when shown as a function of speed. (Figure 22.11) [15] Basically, the machine works as a motor during acceleration from  $n_{\min}$  to  $n_{\max}$  and as a generator from  $n_{\max}$  to  $n_{\min}$  and thus the input instantaneous active power is at times positive or negative.

The average power per cycle represents the average total losses in the IM per cycle  $P_{\text{loss}}$ . By the same token, the average of positive values yields the average input power  $P_{\text{in1m}}$ . So the efficiency  $\eta$  is

$$\eta = 1 - P_{\text{loss}} / P_{\text{in1m}} \quad (22.31)$$

Ref. [15] reports less than 1% difference in efficiency between this method and the direct loading method.

The main demerit of the method is the speed continuous dynamics which gives rise to additional noise and vibration losses in the IM.

If the frequency is changed up and down quickly, the IM speed cannot follow it and thus speed dynamics is avoided. Care must be exercised in this case not to go beyond the d.c. line capacitor threshold voltage during inevitable fast switching from motor to generator mode.

- The PWM dual frequency test [16]

An inverter with an open control architecture is prepared to produce, through PWM, the  $V(t)$  of (22.26) which contains two distinct frequencies.

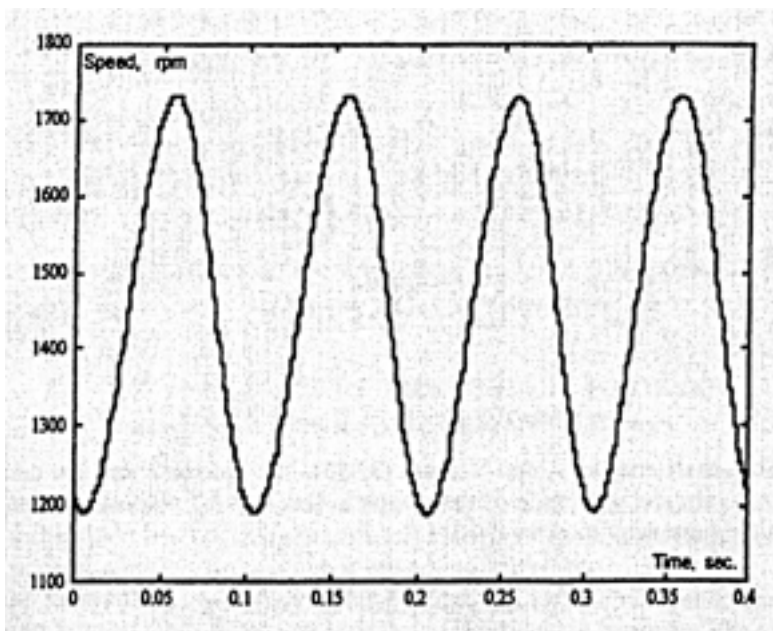
This may be chosen around the rated synchronous speed of the IM:

$$\begin{aligned} f_1 &= f_{in} + \Delta f \\ f_2 &= f_{in} - \Delta f \end{aligned} \tag{22.32}$$

Now  $\Delta f$  is the output of a slow RMS current close loop controller, which will result in the required current load.

Again,  $\Delta f$  is so large that the IM speed oscillates very little, somewhere below the average of the two frequencies. The switching from motoring to generating is now done, by the electromagnetic field, with a frequency equal to the difference between  $f_1$  and  $f_2$ .

To provide for the same core losses, they are considered proportional to voltage squared and superposition is applied.





with  $\Delta f = 5 \times 10^{-2} f_{in}$  from (22.33) and (22.34)  $V_2 = 0.5383 V_{in}$  and  $V_1 = 0.8427 V_{in}$ . A rather complete block diagram of the entire system is shown in Figure 22.12. As evident in Figure 22.12, the whole testing process is “mechanized.” This includes starting the IM with a single frequency voltage, which is ramped slowly until the rated frequency is reached. Then gradually, the dual frequency voltage PWM enters into action and produces the recalculated voltage amplitudes and frequencies. It should be noted that as the load is varied, through the RMS current, not only  $\Delta f$  varies but, also the voltage amplitudes  $V_1$  and  $V_2$  vary according to (22.33)–(22.34). Typical test results are shown in Figure 22.13. [16]

As expected the voltage, flux, and current are modulated. Also the capacitor voltage in the d.c. link varies. When it increases, generating mode is encountered.

The instantaneous active power may be calculated as in the previous method, in synchronous coordinates. This time a few electrical cycles may be averaged but in essence the average of instantaneous active power represents the total losses of the machine. The average of the positive values yields the average input power  $P_{in1m}$ . So the efficiency may be calculated again as in (22.32) for various reference stator RMS currents. Ultimately, the efficiency may be plotted versus output power, which is the difference between input power and losses.

Also temperature measurements may be added either to check the validity of the loss equivalence or, for a given temperature, to calibrate the reference RMS current value.

Temperature measurements [16] seem to confirm the method though verifications on many motors are required for full validation.

It should be emphasized that the IM current has been raised up to 150% and thus the complete efficiency test up to 150% load may be replicated by the PWM dual frequency test.

Notice that the slip frequency is around the rated value and thus the skin effect is not different from the case of direct load testing.

The mixed frequency methods may be used also for the temperature rise tests.

One more indirect method for temperature rise testing, used extensively by a leading manufacturer for the last 40 years for powers up to 21MW, is presented in what follows. [17]

### **22.3 THE TEMPERATURE-RISE TEST VIA FORWARD SHORTCIRCUIT (FSC) METHOD**

For the large power IMs, direct loading to evaluate efficiency and temperature rise under load involves high costs. To avoid direct (shaft) loading, the superposition and equivalent (artificial) loading methods have been used for large powers, non-standard frequency or vertical shaft IMs.

The superposition test relies on the fact that temperature rise due to various losses adds up and thus a few special tests could be used to “simulate” the actual temperature rise in the IM. Scaling of results and the rather impossibility to



consider correctly the stray load losses renders such methods as less practical for IMs.

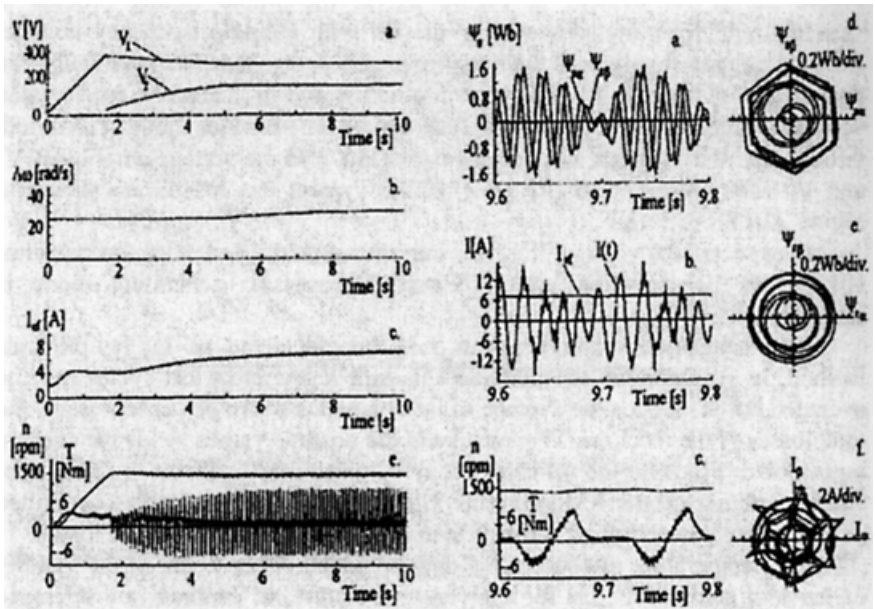


Figure 22.13. Sample test results with the PWM dual frequency method

The forward shortcircuit test (FSC) [17] performs a replica of load testing from the point of view of losses. It is not standardized yet but it uses standard hardware. In essence the tested IM is driven to rated speed slowly by drive motor 1 rated at 10% the rated power of the tested IM (Figure 22.14).

Then a synchronous generator is rotated by drive motor 2, rated at 10% of generator rated power, and then excited to produce low voltages of frequency  $f'_1$ .

$$f'_1 \approx (0.8 - 0.85)f_{in} \quad (22.35)$$

As the generator excitation current increases, so do the generator voltages supplying the tested motor. The tested motor starts operating as a generator feeding active power to the generator, which becomes now a motor (albeit at low power) and the drive motor 2 works as a generator. The frequency of the currents in the rotor of the tested motor  $Sf'_1$  is

$$Sf'_1 = f'_1 - n_p p_1 = -(0.15 - 0.2)f_{in} \quad (22.36)$$

To secure low active power delivery from the tested motor, the slip frequency  $f_2$  is such that the generator operates beyond peak torque slip frequency.

$$|Sf_1'| > f_{2k} \approx \frac{R_2}{2\pi L_{sc}} \quad (22.37)$$

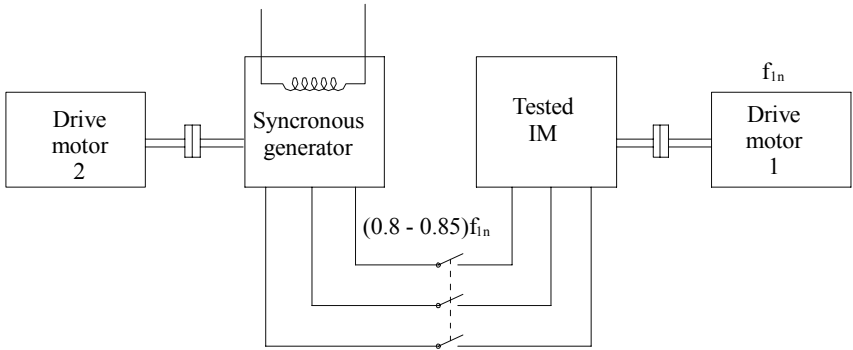


Figure 22.14 Test arrangement for forward shortcircuit test

Consequently, the equivalent impedance of the tested IM is small (large slip values), and, to limit the currents around rated value, the voltage at its terminals has to be low. The synchronous generator excitation current controls the voltage at the tested motor terminals. So low active power is delivered by the IM, albeit at rated current. However, operating beyond the peak torque point may result in instabilities unless the excitation current of the generator is not regulated to “freeze” the operation into a point on the descending torque/speed curve.

As low active powers are handled, the drive motors rating may be as low as 10% of tested motor power. So, in fact, the tested motor experiences rated current but low power (torque) in the generator regime at rather large slip values (0.15–0.25) and low voltage.

The fundamental problem in using this method intended to replicate temperature rise for rated loading is, again, equivalence of losses.

But before that, choosing the generator frequency  $f_1'$  is crucial.

When the test  $f_1'$  frequency decreases the copper losses in the tested motor tend to increase (slip increases) and the required rating of drive motor 1 decreases.

In general, as the slip is rather large the rotor current, for a given stator current, is higher than for rated load conditions (low slip). The difference is generated by the no load (magnetization) current, as the rotor current tends to get closer to stator current with increasing slip.

For low no load current IMs (low number of poles and/or high power), the increase in rotor current in the test with respect to rated power conditions is rather small. Not so for large number of poles and (or) low power IMs.

For  $f_1' = 40$  Hz and  $f_{in} = 50$  Hz the slip frequency ( $Sf_1$ ) = 10 Hz and thus, at least for medium and large power cage rotor IMs, the skin effect is notable in contrast to rated power conditions.

This is an inherent limitation of the method. Let us now discuss the various loss components status in the FST:

- The stator copper losses

As the current is kept at rated value only, the lower skin effect, due to lower frequency  $f_1$  ( $f_1 < f_{1n}$ ), will lead to slightly lower copper losses than in rated load conditions in large power IMs. If the design data are known, with  $f_1$  given, the skin effect resistance coefficient may be calculated for  $f_{1n}$  and  $f_1$  (Chapter 9). Corrections then may be added.

- The fundamental iron losses are definitely lower than for rated conditions as  $f_1$  is

lower than for rated conditions and so is the voltage, even the flux.

To a first approximation, the fundamental core losses are proportional to voltage squared (irrespective of frequency). Also, there are some fundamental core losses in the rotor as  $Sf_1$  is (0.15–0.25)  $f_1$ .

The no load iron losses at rated voltage and frequency conditions may be compared with the low iron losses estimated for the FSC with the difference added by increasing stator current to compensate for the difference in the loss balance.

- Rotor cage fundamental losses

As indicated earlier, the rotor cage losses are larger and dependent on the generator frequency  $f_1$ , but, to calculate their difference, would require the magnetization reactance and the rotor resistance  $R_2$  and leakage reactance  $X_{l2}$  at the rather large frequency  $Sf_1$  (with skin effect considered).

- Stray load losses

A good part of stray load losses occur in the rotor bars. The frequency  $S_v f_1$  of the rotor bar harmonics currents due to stator m.m.f. space harmonics is

$$S_v f_1' = [1 - v(1 - S)]f_1' \quad (22.38)$$

For phase belt harmonics,

$$v = \pm 6K + 1 \quad (22.39)$$

$$\text{with } v = -5, +7, -11, 13 \dots \quad (22.40)$$

If for FST  $f_1 = 40$  Hz and  $S = -0.2375$ , for rated load conditions  $f_{1n} = 50$  Hz and  $S_n = 0.01$ . The rotor current harmonics frequency  $S_v f_1$  are very close in the two cases (287.5/297.5, 306.5/296.5, 584.5/594.5, 603.5/593.5).

So, from this point of view, the two cases are rather equivalent. That is, the FST method is consistent with the direct load method. However, the rotor current harmonics are also influenced by slot permeance harmonics, which are voltage dependent. So in general, the stator m.m.f. caused rotor bar stray losses are smaller for the FST than for rated conditions.

On the other hand, tooth flux pulsation core losses have to be considered. As the saturation is not present (low voltage; low main flux level), the flux pulsation in the tooth, for given currents, tends to be larger in the FST, than for rated load conditions.

The rotor slot harmonic pole pairs  $p_\mu$  is

$$p_{\mu} = KN_r \pm p_1 \quad (22.41)$$

$N_r$  – number of rotor slots,  $p_1$  – IM pole pairs.

The frequency  $f_{\mu}$  of the eddy currents induced in the stator tooth is

$$f'_{\mu} = [p_{\mu}(1-S) + S]f_1 \quad (22.42)$$

It suffices to consider  $K = 1, 2$ .

For the same example as above ( $f'_1 = 40$  Hz,  $S = -0.2375$  and  $f_1 = 50$  Hz,  $S_n = 0.01$ ) and  $K = 1, 2$ , again frequencies  $f_{\mu}$  very close to each other are obtained for the two situations.

So the stator harmonics core losses are equivalent.

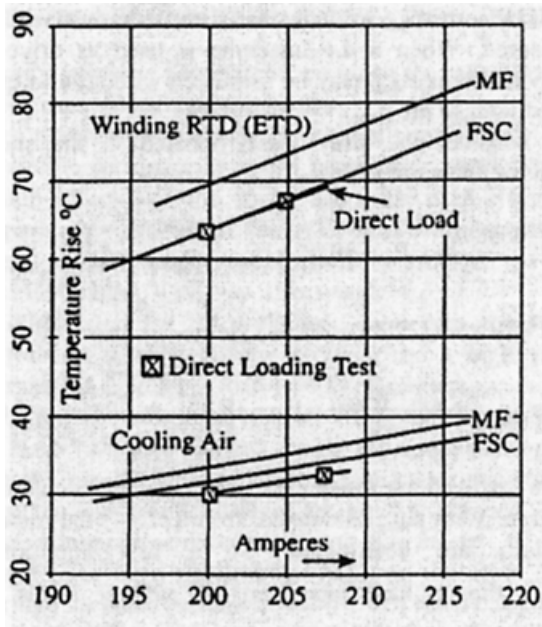


Figure 22.15 Steady state temperature rises with three tests

Still we have to remember that the rotor cage fundamental losses tend to be somewhat larger (because of much larger slip and skin effect) while the fundamental core losses are much lower.

The degree to which these two effects neutralize each other is an issue to be tackled for every machine separately.

Final results related to temperature rise for three tests – FSC, IEEE-112F and the two frequency methods (MF) – are shown on [Figure 22.15](#) for a 1960 kW 4 pole, 50 Hz, 6600 V IM. [17] The FSC and direct loading tests produce very close results while the mixed frequency (MF) method overestimates the temperature by 5–7°C.

The standard mixed-frequency method overestimation of temperature rise (Figure 22.15) is due to lower speed, speed oscillations, and due to higher core losses as the supply frequency voltage  $V_1$  comes in with 100% with  $V_2$  added to it. It appears that the PWM dual frequency test is free from such limitations and would produce reliable results in temperature rise as well as in efficiency tests. Returning to FSC, we notice that if the core losses are neglected (the voltage is reduced), the difference between the generated power  $P_G$  and the shaft input power  $P_{\text{shaft}}$  for the tested motor is

$$P_{\text{shaft}} - P_G = 3I_1^2 R_1 + 3I_2^2 R_2 + p_{\text{stray}} + p_{\text{mec}} + p_{\text{iron}} \left( \frac{V}{V_{\text{ln}}} \right)^2 \quad (22.43)$$

The mechanical losses  $p_{\text{mec}}$  are considered to be known while,  $P_{\text{shaft}}$ ,  $P_G$ ,  $I_1$  have to be measured. When a d.c. machine is used as drive motor 1, its calibration is easy and thus  $P_{\text{shaft}}$  may be estimated without a torque-meter. The iron losses at rated voltage are  $p_{\text{iron}}$ .

Now if we consider the stray losses located in the stator, then the electromagnetic power concept yields

$$\frac{3I_2^2 R_2}{|S|} = P_G + 3R_1 I_1^2 + p_{\text{stray}} + p_{\text{iron}} \left( \frac{V}{V_{\text{ln}}} \right)^2 \quad (22.44)$$

Eliminating  $I_2^2 R_2$  from (22.43)–(22.44) yields

$$p_{\text{stray}} = \frac{(P_{\text{shaft}} - p_{\text{mec}})}{1 + |S|} - P_G - 3R_1 I_1^2 - p_{\text{iron}} \left( \frac{V}{V_{\text{ln}}} \right)^2 \quad (22.45)$$

As all powers involved are rather small, the error of calculating  $p_{\text{stray}}$  is not expected to be large. With  $p_{\text{stray}}$  calculated, from (22.44) the rotor fundamental cage losses  $3R_2 I_2^2$  are determined. To a first approximation, with  $I_2 \approx \sqrt{I_1^2 - I_m^2}$  ( $I_m$  – the no load current at low voltage levels that is under unsaturated conditions), the rotor resistance  $R_2$  at  $Sf_1$  frequency may be determined as a bonus. Comparing results on parameter calculation from FSC and IEEE–112F shows acceptable correlation for the 1960 kW IM investigated in Reference 17

The level of voltage at IM terminals during FSC is around 20% rated voltage. A few remarks on FSC are in order

- The loss distribution is changed with more losses in the rotor and less in the stator.
- The skin effect is rather notable in the rotor.
- It is possible to replace the drive motor 2 and the a.c. generator by a bi-directional power PWM converter sized around 20% the rated power of the tested IM (Figure 22.16).

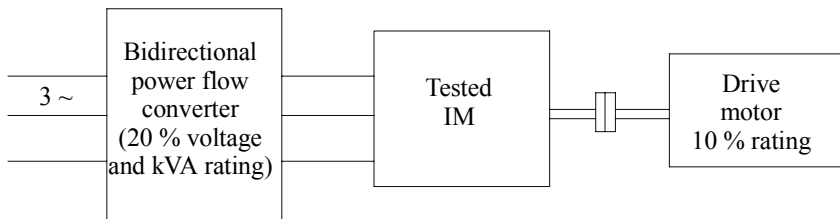


Figure 22.16 The forward shortcircuit (FSC) test with PWM converter

The control options, for adequate loss equivalence and stable operation, are improved by the presence and the capabilities of the PWM converter.

## 22.4 PARAMETER ESTIMATION TESTS

By induction machine parameters, we mean the resistances and inductances in the adapted circuit model and the inertia.

There are many ways we may develop a circuit model, which duplicates better or worse the performance of IM in different operation modes. Magnetic saturation, skin effect and the space and time m.m.f. harmonics make the problem of parameter estimation very difficult. Traditionally, the no load and stall rotor tests at rated frequency are used to estimate the parameters of the single cage circuit model, with skin effect neglected.

On the other hand for IM with deep rotor bars, or double cage rotors, shortcircuit tests at rated frequency produce too a high value for the rotor resistance for the IM running at load and rated frequency.

Standstill frequency response tests are recommended for such cases.

Still the large values of currents during IM direct starting at the power grid produce saturation in the leakage path of both stator and rotor magnetic field. Consequently, the leakage inductances are notably reduced. This effect adds to the rotor leakage inductance reduction due to skin effect.

The closed rotor cage slot leakage field, on the other hand, saturates the upper iron bridge for rotor currents above 10% of rated current.

This means that for such a case the rotor leakage inductance does not vary essentially at currents above 10% rated current, that is during starting or on load conditions.

For variable speed IM operation, the level of the airgap flux varies notably. Consequently, the magnetization inductance varies also. The apparent way out of these difficulties may seem to use the traditional single-rotor circuit model with all parameters as variables (Figure 22.17) with primary ( $f_1$ ), rotor slip frequency  $Sf_1$ , and stator, rotor and magnetization currents ( $I_1$ ,  $I_2$ ,  $I_m$ ). The trouble is that most parameters depend on more than one variable and thus such an equivalent circuit is hardly practical (Figure 22.17). Also, if space and time harmonics are to be considered, the general equivalent circuit becomes all but manageable.

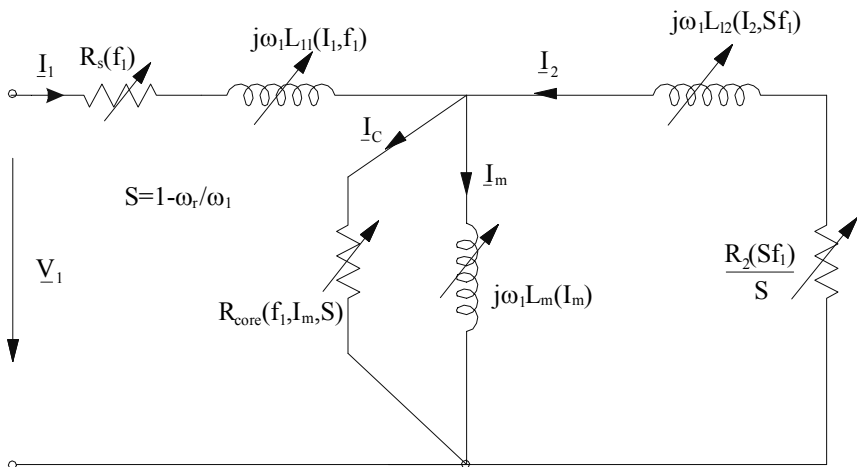


Figure 22.17 Fundamental single cage rotor equivalent circuit with variable parameters

For the space harmonics, the slip  $S$  changes to  $S_v$  while for time harmonics both  $f_k$  and  $S_k$  vary with the harmonics order  $K$  (see chapter 10 on harmonics). We should remember that, while spinning the rotor, slip frequency  $Sf_1$  is different from stator frequency  $f_1$  and thus is full reality only tests done in such conditions match the performance. Also, FEM is now approaching enough maturity to be useful in the computation of all parameters in any conditions of operation.

Finally, for IM control for drives a rather simplified equivalent circuit (model) with some parameter variation is practical. This is how simplified solutions for IM parameter estimation off line and on line evolved.

Among them, in fact only the no load and shortcircuit single frequency test derived parameter estimation method is standardized worldwide. The two frequency and the frequency response standstill tests have gained some momentum lately. Modified regression methods to estimate double cage model parameters based on acquiring data over a direct slow starting period or directly from transients during a fast start, have been developed recently. Finally for variable speed drives commissioning and their adaptive control, off and on line simplified methods to determine some parameters detuning are currently used. In what follows, a synthesis of these trends is presented.

#### 22.4.1 Parameter calculation from no-load and standstill tests

The no-load motor test – paragraph 22.1.1. – is hereby used to calculate the electric parameters appearing in the equivalent circuit for zero rotor currents (Figure 22.18).

With voltage  $V_1$ , current  $I_{10}$ , and power  $P_0$  measured, we may calculate two parameters:

$$R_1 + R_{cs} = \frac{(P_0 - p_{mec})}{3I_{10}^2} \quad (22.46)$$

$$L_{11} + L_m = \frac{1}{\omega_1} \sqrt{\left(\frac{V_1}{I_{10}}\right)^2 - (R_1 + R_{cs})^2}$$

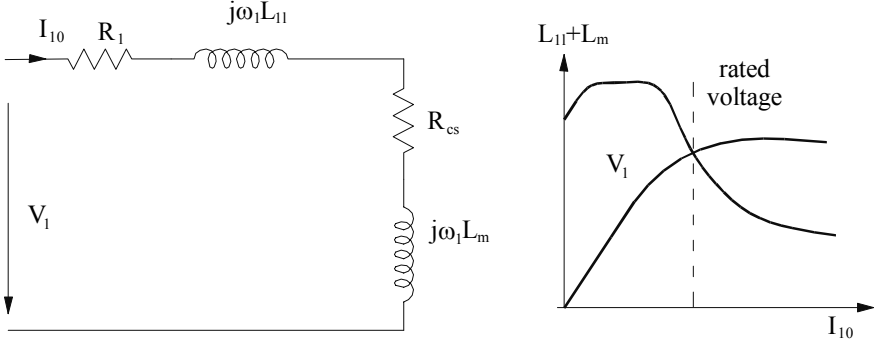


Figure 22.18 Fundamental no load equivalent circuit

The stator phase resistances may be d.c. measured or, if possible, in a.c. without the rotor in place. In this case, if a small three phase voltage is applied, the machine will show a very small  $R_{cs}$  (core losses are very small). The calculated phase inductance slightly over estimates the stator leakage inductance.

The overestimation is related to the magnetic energy in the air left by the missing rotor.

The equivalent inductance of the “air” is, in fact, corresponding to the magnetization induction of an equivalent airgap of  $\tau/\pi$  ( $\tau$  the pole pitch).

$$L_g = (L_m)_{unsat} \cdot \frac{g\pi}{\tau} \quad (22.47)$$

Suppose we do these measurements (without the rotor) and obtain from power, voltage, and current measurements,

$$R_{1\sim} = \frac{P_{3\sim}}{3I_{10\sim}^2} \quad (22.48)$$

$$L_{11} + L_g = \frac{1}{\omega_1} \sqrt{\left(\frac{V_{1\sim}}{I_{10\sim}}\right)^2 - R_{1\sim}^2}$$

From the no-load test at low voltage, the unsaturated value of  $L_m$  is obtained. Then from, (22.47)  $L_g$  is easily calculated, provided the airgap is known. Finally from (22.48)  $L_{11}$  is found with  $L_g$  already calculated.

In general, however, the stator leakage inductance is not considered separable from the no load and shortcircuit tests.



The shortcircuit (standstill or stall) tests, on the other hand, are performed at low voltage with three phase or single-phase supply (paragraph 22.1.4), and  $R_{sc}$  and  $X_{sc} = \omega_1 L_{sc}$  are calculated from (22.9)–(22.10) and averaged over a few current values up to about rated current. When voltage versus current is plotted, the curve tends to be linear but, for the closed-slot rotor, it does not converge into the origin (Figure 22.19).

The residual voltage  $E_{bridge}$ , corresponding to the rapidly saturating iron bridge above the rotor closed slot, may be added to the equivalent circuit in the rotor part. All the other components of the rotor leakage inductance occur in  $L_{l2}$ . Such an approximation appears to be practical for small induction machines.

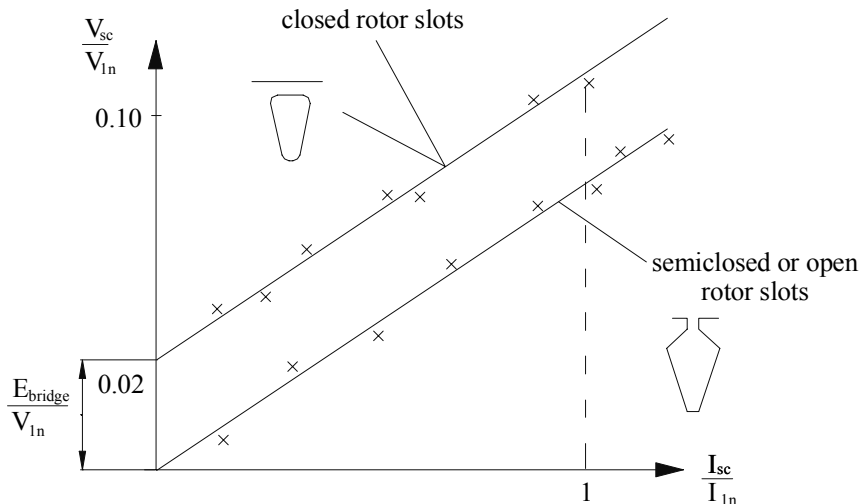


Figure 22.19 Typical voltage/current curves at standstill

Now, with  $R_1$  known,  $R_{2sc}$  may be calculated from (22.9)–(22.10). That is, the rotor resistance for the rated frequency, with the skin effect accounted for.

The stator and rotor leakage inductances at shortcircuit come together in  $X_{sc} = (L_{l1} + L_{l2sc})\omega_1$ . Unless  $L_{l1}$  is measured as indicated above or its design value is known,  $L_{l1}$  is taken as  $L_{l1} \approx L_{sc}/2$ .

The no load and shortcircuit tests for the wound rotor IM are straightforward as both the stator and rotor currents may be measured directly. Also the tests may be run with stator or rotor energized. Even both stator and rotor circuits may be energized. Changing the amplitude and phase of rotor voltage until the airgap field is zero, leads to a complete separation of the two circuits. The same tests have been proved acceptable to derive torque/current load characteristics for superhigh speed PWM converter fed IMs. [18] The current is limited in this case by the converter rating. Neglecting the magnetization inductance in the equivalent circuit at standstill ( $S = 1$ ) – Figure 22.20 – is valid as long as  $\omega_1 L_m \gg R_{2sc}$ .

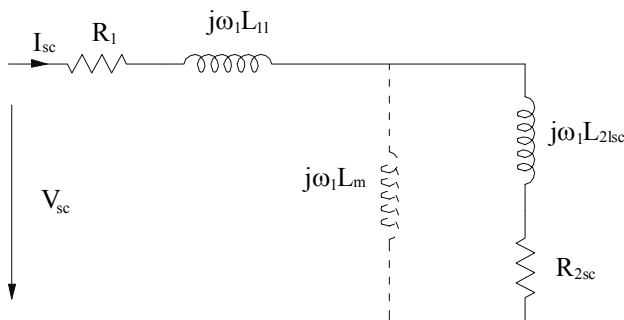


Figure 22.20 Equivalent circuit at standstill

It is known that at full voltage, the airgap flux is only about (50–60%) of its no load value due to the large voltage drop along the stator resistance  $R_1$  and leakage reactance  $\omega_1 L_{l1}$ . So, in any case, the main magnetic circuit is nonsaturated. That is to say that the nonsaturated value of  $L_m - (L_m)_{\text{unsat}}$  – has to be introduced in the equivalent circuit at low frequencies when  $\omega_1 L_m$  is not much larger than  $R'_{2sc}$ . If the  $L_m$  branch circuit is left out in low frequency standstill tests, the errors become unacceptable.

The presence of skin effect suggests that the single rotor circuit model is not sufficient to accommodate a wide spectrum of operation modes from standstill to no load.

The first easy step is to perform one more standstill test, but at low frequency: less than 5% of rated frequency. This is how the so-called two-frequency standstill test was born.

#### 22.4.2 The two frequency standstill test

The second shortcircuit test, at low frequency, requires a low frequency power source. The PWM converter makes a good (easily available) low frequency power supply.

This time the equivalent circuit contains basically two loops (Figure 22.21)

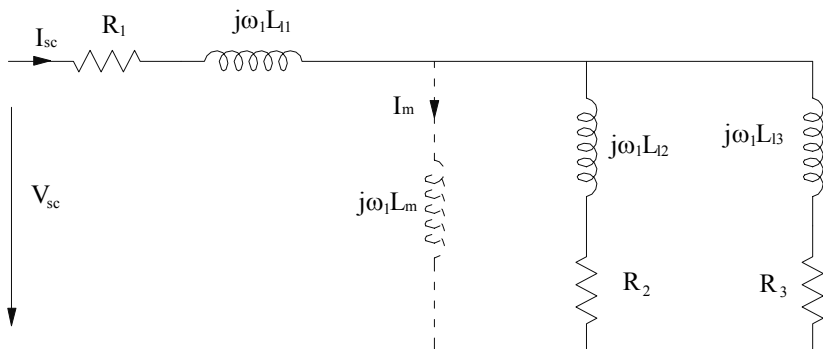


Figure 22.21 Dual rotor – loop equivalent circuit at standstill

To calculate the four rotor parameters  $R_2$ ,  $L_{12}$ ,  $R_3$ ,  $L_{13}$ , it is assumed that at rated frequency  $f_{1n}(\omega_{1n})$  only the second (starting) rotor loop exists. That is, only  $L_{13}$  and  $R_3$  remain as unknowns and the  $L_m$  branch may be eliminated. At the second frequency  $f_1 \leq 0.05f_{1n}$ , only the first (working) cage acts. Thus, only  $R_2$ ,  $L_{12}$  are to be calculated with the  $L_m$  branch accounted for.  $L_m$  is taken as the nonsaturated value of  $L_m$  from the no load test. An iterative procedure may be used with  $L_m(I_m)$  from the no load test as the main circuit may get saturated unless  $V_{sc}/V_{1n} < f_1/f_{1n}$ .

The rotor parameters of rotor loops are constant. One dominates the performance at large slip frequencies  $Sf_1$  ( $R_3$ ,  $L_{13}$ ) and the other one prevails at low slip frequency.

The double-cage rotor IM may be assimilated with this case. However, the two cages measured parameters may not overlap with the actual double-cage parameters.

As of now, with the above parameter estimation methods, skin effect has been accounted for approximately. Also by the  $L_m(I_m)$  function, the magnetic saturation of the main circuit is accounted for.

Still, the leakage saturation at high currents has not been present in tests. Consequently, calculating the torque, current, power factor from  $S = 1$  to  $S_n$  ( $S_n$  – rated slip) with the above parameters is bound to produce large errors. And so it does!

The dissatisfaction of IM users with this situation led to quite a few simplified attempts to calculate the parameters, basically  $R_2$ ,  $R_{2sc}$ ,  $L_{12}$ ,  $L_{12sc}$ , based on catalogue data to fit both the starting current, and torque and the rated current and power factor for given efficiency, rated speed and torque. These approaches may be dubbed as catalogue based methods.

### 22.4.3 Parameters from catalogue data

Let us consider that the starting current and torque  $I_s$ ,  $T_{es}$ , rated speed, frequency, current, power factor, efficiency and rated power are all catalogue data for the IM. It is well understood that the logic solution to parameter estimation for two extreme situations – start and full load – is to use the equivalent circuit. First from the no-load current  $I_0$ ,

$$I_0 \approx \frac{V_{1n}}{\omega_1 L_m (1 + L_{11} / L_m)} \quad (22.49)$$

We obtain an approximate value of  $L_m$ . For start  $L_{11}/L_m = 0.02-0.05$ . At the end of the estimation process,  $L_{11}$  is obtained.

Then we may return with a corrected value of  $L_{11}/L_m$  to (22.49) and perform the whole process again until sufficient convergence is reached.

This saturated value is valid for rated load calculations.

From the starting torque,

$$T_{es} = \frac{3R_{2sc} \cdot I_{2s}^2 \cdot p_l}{\omega_l} \quad (22.50)$$

$$I_{2s} \approx \sqrt{I_s^2 - I_0^2}$$

From (22.50),  $R_{2sc}$  may be determined. The equivalent circuit at start yields

$$(L_{lsc} + L_{l2sc}) = \frac{1}{\omega_l} \sqrt{\left(\frac{V_{ln}}{I_s}\right)^2 - (R_l + R_{2sc})^2} \quad (22.51)$$

The stator resistance  $R_l$  may be d.c. measured and thus  $(L_{lsc} + L_{l2sc})$  as influenced by skin effect and leakage saturation, is determined.

On full load, the mechanical losses are assigned an initial value  $p_{mec} = 0.01P_n$ .

Thus the electromagnetic torque is equal to the shaft torque plus the mechanical loss torque.

$$T_{en} = p_l \left(1 + \frac{p_{mec}}{P_n}\right) \frac{P_n}{2\pi f_l (1 - S_n)} = \frac{3R_2 I_{2n}^2}{S_n \omega_l} p_l \quad (22.52)$$

$$I_{2n}^2 \approx \sqrt{I_{ln}^2 - I_0^2}$$

With  $R_2$  found easily from (22.52), we may write the balance of powers at rated slip,

$$Q_l = 3V_{ln} I_{ln} \sin \phi_{ln} = 3\omega_l L_{l1} I_{ln}^2 + 3\omega_l L_{l2} I_{2n}^2 + 3L_m \omega_l \cdot I_0^2 \quad (22.53)$$

$$P_l = 3V_{ln} I_{ln} \cos \phi_{ln} = 3R_l I_{ln}^2 + p_{iron} + p_{stray} + 3 \frac{R_2 I_{2n}^2}{S_n} \quad (22.54)$$

$$P_n \left(1 - \frac{1}{\eta}\right) = 3R_l I_{ln}^2 + p_{iron} + p_{stray} + 3R_2 I_{2n}^2 + p_{mec} \quad (22.55)$$

There are 5 unknowns –  $L_{l1}$ ,  $L_{l2}$ ,  $p_{iron}$ ,  $p_{stray}$ ,  $p_{mec}$  – in Equations (21.53-21.55).

However from (21.54),  $p_{iron} + p_{stray}$  can be calculated directly. They may be introduced in (21.55) to calculate mechanical loss  $p_{mec}$ . If they differ from the initial value of 1% of rated power  $P_n$ , their value is corrected in (21.52) and  $p_{core} + p_{stray}$  and  $p_{mec}$  are recalculated again until sufficient convergence in  $p_{mec}$  is obtained.

We thus remain with Equation (21.53) where in fact we do have two unknowns: the stator and rotor leakage inductances. They may not be separated and thus we may consider them equal to each other:  $L_{l1} = L_{l2}$ .

This way

$$L_{11} = L_{12} = \frac{(3V_{1n}I_m \sin \phi_{1n} - 3\omega_1 L_m I_0^2)}{3\omega_1 (2I_{1n}^2 - I_0^2)} \quad (22.56)$$

Now going back to (22.51), if the IM has open stator slots, there will be no leakage flux path saturation in the stator and thus  $L_{11sc} = L_{11} = L_{12}$ . Consequently, from (22.51),  $L_{12sc}$  is found. With semiclosed stator slots, both stator and rotor leakage flux paths may saturate. So, we might as well consider  $L_{11sc} = L_{12sc}$ . Again  $L_{12sc} = L_{11sc}$  is found from (22.51).

Attention has to be paid to (21.56) as computation precision is important because most of the reactive power is “spent” in the airgap, in  $L_m$ . In general,  $2L_{11} > L_{11sc} + L_{12sc}$ . If  $2L_{11} \leq L_{11sc} + L_{12sc}$  is obtained the skin and saturation effects are mild and the shortcircuit values hold as good for all slip values.

An approximate dependence of rotor resistance and leakage inductances with slip, between  $S = 1$  and  $S = S_n$  may be admitted

$$R_2(S) = R_2 + (R_{2sc} - R_2) \sqrt{\frac{(S - S_n)}{1 - S_n}}; S \geq S_n \quad (22.57)$$

$$L_{12}(S) = L_{12} - (L_{12} - L_{12sc}) \frac{S - S_n}{1 - S_n}; S \geq S_n \quad (22.58)$$

$$L_{11}(S) = L_{11} - (L_{11} - L_{11sc}) \frac{S - S_n}{1 - S_n}; S \geq S_n \quad (22.59)$$

This way the parameters dependence on slip is known, so torque, current, power factor versus slip may be calculated. Alternatively, we may consider that the rotor parameters at standstill  $R_{2sc}$ ,  $L_{12sc}$  represent the starting cage and  $R_{21}$ ,  $L_{12}$  the working cage. Such an attitude may prove practical for the investigation of IM transients and control.

It should be emphasized that catalogue data methods provide for complete agreement between the circuit model and the actual machine performance only at standstill and at full load. There is no guarantee that the rotor parameters vary linearly or as in (21.57) with slip for all rotor (stator) slot geometries.

In other words, the agreement of the torque and current versus slip curves for all slips is not guaranteed to any definable error.

From the need to secure good agreement between the circuit model and the real machine, for ever wider operation modes, more sophisticated methods have been introduced.

We introduce here two of the most representative ones: the standstill frequency response method (SSFR) and the step-wise regression general method.

### 22.4.4 Standstill frequency response method

Standstill tests, especially when a single phase ac supply is used (zero torque), require less manpower, equipment and electrical energy to perform than running tests. To replicate the machine performance with different rotor slip frequency  $Sf_1$ , tests at standstill with a variable voltage and frequency supply have been proposed [19-20]. Dubbed as SSFR, such tests have been first used for synchronous for generator parameter estimation.

In essence, the IM is fed with variable voltage and frequency from a PWM converter supply.

Single phase supply is preferred as the torque is zero and thus no mechanical fixture to stall the rotor is required. However, as three phase PWM converters are available up to high powers and the tests are performed at low currents (less than 10% of rated current), three phase testing seems more practical (Figure 22.22)

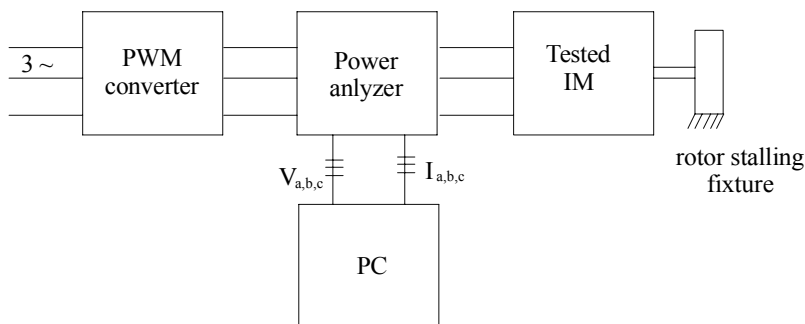


Figure 22.22 Standstill frequency test arrangement

The test provides for the consideration of skin effect, which is frequency dependent. However, as the current is low, to avoid IM overheating, the SSFR test methods should not be expected to produce good agreement with the real machine in the torque/slip or current/slip curves at rated voltage. This is so because the leakage saturation effect of reducing the leakage inductances is not considered. A way out of this difficulty is to adopt a certain functional for stator leakage inductance variation with stator current.

Also at low frequencies, 0.01 Hz, 2–3 periods acquisition time means 200–300 seconds. So, in general, acquiring data for 50–100 frequencies is time prohibitive. The voltages and currents are acquired and their fundamental wave amplitudes and ratio of amplitudes and phase lag are calculated.

A two (even three) rotor loop circuit model may be adopted (Figure 22.23).

The operational inductance  $L(p)$  for a double cage model (see Chapter 13 on transients) in rotor coordinates, is:

$$L(p) = (L_{11} + L_m) \frac{(1 + pT'_1)(1 + pT''_1)}{(1 + pT'_0)(1 + pT''_0)} \quad (22.60)$$

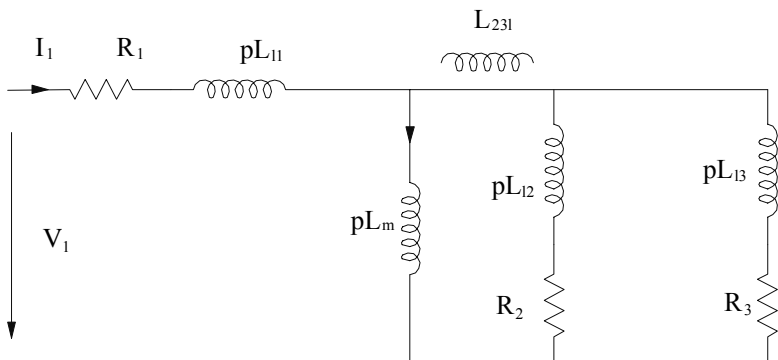


Figure 22.23 Double cage operational equivalent circuit at standstill

The equivalent circuit does not show the mutual leakage inductance between the two virtual cages ( $L_{123}$ ) because it can not be measured directly.

The time constants in (22.60) are related to the equivalent circuit parameters as follows

$$T' = \frac{1}{R_3} \left[ L_{31} + \frac{L_m \cdot L_{11}}{L_m + L_{11}} \right] \quad (22.61)$$

$$T'' = \frac{1}{R_2} \left[ L_{12} + \frac{L_m L_{11} L_{13}}{L_{13} L_m + L_{13} L_{11} + L_m L_{11}} \right] \quad (22.62)$$

$$T'_0 = \frac{1}{R_3} (L_m + L_{11} + L_{13}) \quad (22.63)$$

$$T''_0 = \frac{1}{R_2} \left[ L_{12} + \frac{L_m L_{13}}{L_m + L_{13}} \right] \quad (22.64)$$

The unsaturated value of the no load inductance  $L_s = L_{11} + L_m$  has to be determined separately from the no load test at lower than rated voltage.

Also the stator leakage inductance value has to be known from design or is being assigned a “reasonable” value of 2 – 5% of no load inductance  $L_s$ .

The measurements will result in impedance estimation for every frequency test performed.

$$\underline{Z}_{sc}(\omega) = \frac{V_{sc}}{I_{sc}} = R_{sc}(\omega) + jX_{sc}(\omega) \quad (22.65)$$

When the IM is single-phase supply, fed between 2 phases, a factor of 2 occurs in the denominator of (22.65).

We may now calculate  $L(j\omega)$  from (22.65).

$$L(j\omega) = \frac{(Z_{sc} - R_1)}{j\omega_1} = \text{Re}(\omega) + j \text{Im}(\omega) \quad (22.66)$$

Equating the real and imaginary parts in (22.60) and (22.66), we obtain

$$\text{Re}(\omega) = \omega L_s + \omega \text{Im}(\omega) (T'_0 + T''_0) - \omega^3 L_s T' T'' + \omega^2 \text{Re}(\omega) \cdot T'_0 \cdot T''_0 \quad (22.67)$$

$$\text{Im}(\omega) = \omega^2 L_s (T' + T'') - \omega \text{Re}(\omega) (T'_0 + T''_0) - \omega^2 \text{Im}(\omega) T'_0 \cdot T''_0 \quad (22.68)$$

A new set of unknowns, instead of  $T'$ ,  $T''$ ,  $T'_0$ ,  $T''_0$ , is now defined. [21]

$$\begin{aligned} X_1 &= L_s \\ X_2 &= T'_0 + T''_0 \\ X_3 &= L_s T' T'' \\ X_4 &= T'_0 T''_0 \\ X_5 &= L_s (T' + T'') \end{aligned} \quad (22.69)$$

$L_s$  may be discarded by introducing the known value of the non-saturated no load inductance.

A linear system is obtained.

In matrix form,

$$|Z_i| = |H_i| |X|; i = 1, 2, \dots, m \quad (22.70)$$

with

$$|Z_i| = \begin{bmatrix} \text{Re}(\omega) & \text{Im}(\omega) \end{bmatrix}_i^T \quad (22.71)$$

$$|X| = [X_1, X_2, X_3, X_4, X_5] \quad (22.72)$$

$$|H_i| = \begin{bmatrix} \omega & \omega \text{Im}(\omega) & -\omega^3 & \omega^2 \text{Re}(\omega) & 0 \\ 0 & -\omega \text{Re}(\omega) & 0 & \omega^2 \text{Im}(\omega) & \omega^2 \end{bmatrix} \quad (22.73)$$

The measurements are taken for  $m \geq 5$  distinct frequencies.

As the number of equations is larger than the number of unknowns, various approximation methods may be used to solve the problem.

The solution of (22.70) in the least squared error method sense is

$$|X| = [H]^T \cdot [H]^{-1} \cdot [H]^T \cdot [Z] \quad (22.74)$$

Once the variable vector  $|X|$  is known, the actual unknowns  $L_s$ ,  $T'$ ,  $T''$ ,  $T'_0$ ,  $T''_0$  are



$$L_s = X_1$$

$$T', T'' = \frac{1}{2X_1} \left( X_5 \pm \sqrt{X_5^2 - 4X_3X_1} \right) \quad (22.75)$$

$$T_0', T_0'' = \frac{1}{2} \left( X_2 \pm \sqrt{X_2^2 - 4X_4} \right) \quad (22.76)$$

The least squared error method has been successfully applied to an 1288 kW IM in Ref. [21].

Typical direct impedance results are shown on [Figure 22.24](#). [21]

Comparative results [20,21] of parameter values estimated with Bode diagram and, respectively, by the least squared method, for the same machine, are given in [Table 22.3](#). [21]

Good agreement between the two methods of solution is evident.

However, as Ref. [20] shows, when applying these parameters to the torque/slip or current/slip curves at rated load, the discrepancies in torque and current at high values of slip are inadmissible. This may be explained by the neglect of leakage saturation, which in the real machine with higher starting torque and current is notable.

**Table 22.3 Comparative parameter estimates for an 1288 kW IM**

| Parameter    | Bode diagram [10]     | Least squared method [21] |
|--------------|-----------------------|---------------------------|
| $I_s$ p.u    | 5.3                   | 5.299                     |
| $T'$ sec.    | $88 \times 10^{-3}$   | $87.99 \times 10^{-3}$    |
| $T''$ sec.   | $3.6 \times 10^{-3}$  | $3.6 \times 10^{-3}$      |
| $T_0'$ sec.  | 2.0                   | 2.00                      |
| $T_0''$ sec. | $4.55 \times 10^{-3}$ | 4.55                      |

As always in such cases, an empirical expression of the stator leakage inductance, which decreases with stator current, may be introduced to improve the model performance at high currents.

Still, there seem to be some notable discrepancies in the breaking torque region, which hold in there even after correcting the stator leakage inductance for leakage saturation. A three-circuit cage seems necessary in this particular case.

This raises the question of when the SSFR parameter estimation may be used with guaranteed success. It appears that only for small periodic load perturbations around rated speed or for estimating the machine behaviour with respect to high frequency time harmonics so typical in PWM converter-fed IMs.

Step response methods to different levels of voltage and current peak values may be the response to the leakage saturation at high currents inclusion into the IM models.

Short voltage pulses may be injected in the IM at standstill between two phases until the current peaks to a threshold, predetermined value. A pair of

positive and negative voltage pulses may be applied to eliminate the influence of hysteresis.

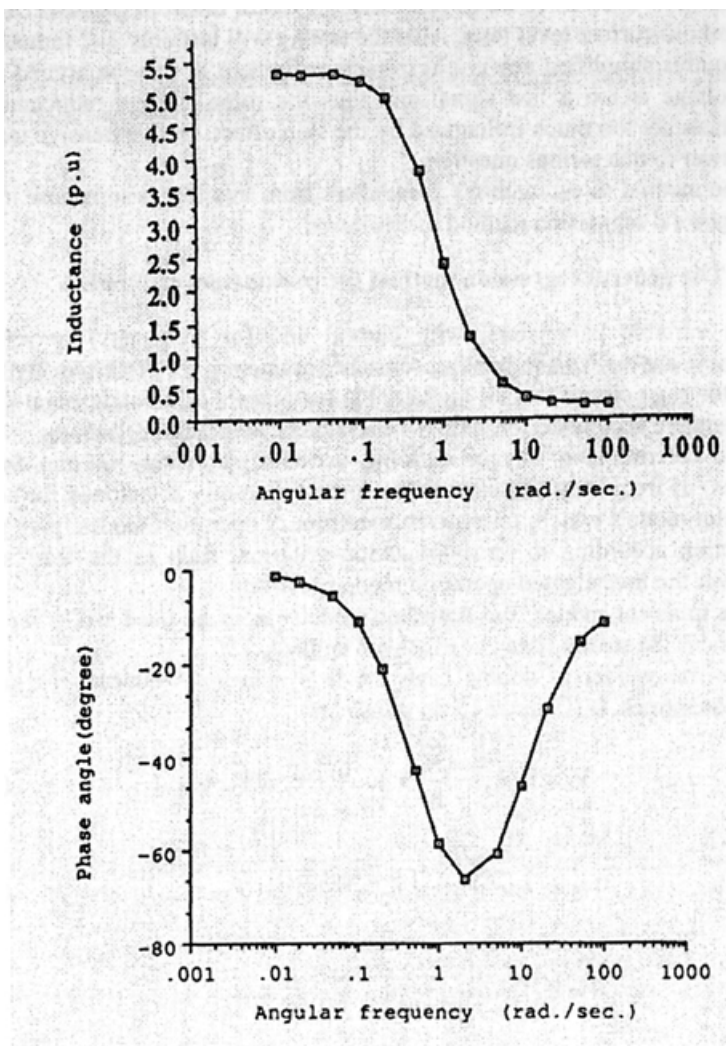


Figure 22.24 Standstill frequency response of an 1288 kW IM

The current and voltage are acquired and processed using the same operational inductance (22.60).

$$v(p) = 2 \cdot i(p) [R_1 + pL(p)] \quad (22.77)$$

The two cage model may be used, with its parameters determined to fit the current/time transient experimental curve via an approximation method (least squared method, etc.).

The voltage signals will stop, via a fast static power switch, at values of current up to the peak starting current to explore fully the leakage saturation phenomenon. The pulses are so short that the motor does not get warm after 10–15 threshold current level tests. Also the testing will last only 1–2 minutes in all and requires simplified power electronics equipment. It may be argued that the step voltage is too a fast signal and thus the rotor leakage inductances and resistances are too much influenced by the skin effect. Only extensive tests hold the answer to this serious question.

The method to estimate the parameters from this test is somehow included in the general regression method that follows.

### 22.4.5 The general regression method for parameters estimation

As variable parameters, with current and (or frequency) or rotor slip frequency, are difficult to handle, constant parameters are preferred. By using a two rotor cage circuit model, the variable parameters of a single cage with slip frequency are accounted for. Not so for magnetic saturation effects.

The deterministic way of defining variable parameters for the complete spectrum of frequency and current has to be apparently abandoned for constant parameter models which, for a given spectrum of operation modes, give general satisfaction according to an optimization criterion, such as the least squared error over the investigated operation mode spectrum.

For transient modes, the transient model has to be used while for steady state modes the steady state equations are applied.

The d.q. model of double cage IM in synchronous coordinates in space phasor quantities, is (Chapter 13 on transients)

$$V_1 = R_1 \bar{I}_1 + \frac{d\bar{\Psi}_1}{dt} + j\omega_1 \bar{\Psi}_1; \bar{\Psi}_1 = \bar{\Psi}_m + L_{11} \bar{I}_1 \quad (22.78)$$

$$0 = R_2 \bar{I}_2 + \frac{d\bar{\Psi}_2}{dt} + jS\omega_1 \bar{\Psi}_2; \bar{\Psi}_2 = \bar{\Psi}_m + L_{12} \bar{I}_2 \quad (22.79)$$

$$0 = R_3 \bar{I}_3 + \frac{d\bar{\Psi}_3}{dt} + jS\omega_1 \bar{\Psi}_3; \bar{\Psi}_3 = \bar{\Psi}_m + L_{13} \bar{I}_3 \quad (22.80)$$

$$\bar{\Psi}_m = L_m (\bar{I}_1 + \bar{I}_2 + \bar{I}_3); \bar{I}_2 = I_{d2} + jI_{q2}; \bar{I}_3 = I_{d3} + jI_{q3} \quad (22.81)$$

$$\frac{J}{p_1} \frac{d\omega_r}{dt} = T_e - T_{load}; T_e = \frac{3}{2} p_1 L_m \cdot \text{Imag} \left[ \bar{I}_1 \cdot \left( \bar{I}_2^* + \bar{I}_3^* \right) \right] \quad (22.82)$$

For sinusoidal voltages

$$\bar{V}_1 = V\sqrt{2} \cos \gamma - jV\sqrt{2} \sin \gamma; \bar{I}_1 = I_{d1} + jI_{q1} \quad (22.83)$$

Equation (22.78)–(22.82) represent a seven order system with 7 variables ( $I_{d1}, I_{q1}, I_{d2}, I_{q2}, I_{d3}, I_{q3}, \omega_r$ ).

For steady state  $d/dt = 0$  in Equations (22.78)–(22.80) – synchronous coordinates.

When the IM is fed from a non-sinusoidal voltage source,  $\bar{V}_s$  has to be calculated according to Park transformation.

$$\bar{V}_s = \frac{2}{3} \left[ V_a(t) + V_b(t)e^{j\frac{2\pi}{3}} + V_c(t)e^{-j\frac{2\pi}{3}} \right] e^{-j\theta_1} \quad (22.84)$$

$$\frac{d\theta_1}{dt} = \omega_1 \quad (22.85)$$

Equation (22.85) represents the case when the fundamental frequency varies during the parameter estimation testing process.

For steady state (22.79)–(22.81) may be solved for torque  $T_e$ , stator current,  $I_1$ , and input power  $P_1$ . [22]

$$T_e(s) = 3V^2 \cdot \frac{P_1}{\omega_1} \cdot \frac{R_r}{S(A_2 + B_2)} \quad (22.86)$$

$$I_1(s) = V \sqrt{\frac{C^2 + D^2}{A^2 + B^2}} \quad (22.87)$$

$$P(s) = 3V^2 \frac{AC - BD}{A^2 + B^2} \quad (22.88)$$

$$R_r = \frac{R_2 R_3 (R_2 + R_3) + (R_2 X_{13}^2 + R_3 X_{12}^2) S^2}{(R_2 + R_3)^2 + (X_{12} + X_{13})^2 S^2} \quad (22.89)$$

$$X_{rl} = \frac{(X_{12} X_{13} (X_{12} + X_{13}) S^2 + R_2^2 X_{13} + R_3^2 X_{12})}{(R_2 + R_3)^2 + (X_{12} + X_{13})^2 S^2} \quad (22.90)$$

$$C = 1 + X_{rl} / X_m; D = R_r / (S X_m); E = 1 + X_{ll} / X_m \quad (22.91)$$

$$A = R_1 C + \frac{R_r}{S} E; B = X_{ll} + X_{rl} \cdot E - R_1 D \quad (22.92)$$

$$X_i = \omega_1 L_i \quad (22.93)$$

The steady state, and transient models have as inputs the phase voltage RMS value  $V$ , its phase angle  $\gamma$ ,  $p_1$  pole pairs, and frequency  $\omega_1$ .

For steady state, the slip value is also a given. The stator resistance  $R_1$ , stator leakage and magnetization inductances,  $L_{l1}$  and  $L_m$ , are to be known apriori.

The stator current or torque or power under steady state, are measured.

Using torque, or current or input power measured functions of slip (from  $S_1$  to  $S_2$ ), the rotor double cage parameters that match the measured curve have to be found. This is only an example for steady state. In a similar way, the observation vector may be changed for transients into stator currents d-q components versus time:  $I_{d1}(t)$ ,  $I_{q1}(t)$ . For a number of time steps  $t_1, \dots, t_n$ , the two currents are calculated from the measured stator current after transformation through Park transformation (22.87).

To summarize, the parameters vector  $|X|$ ,

$$|X| = |R_2, R_3, L_{l2}, L_{l3}|^T \quad (22.94)$$

may be estimated from a row  $|Y|$  of test data at steady state or transients,

$$|Y| = [Y(S_1), Y(S_2), \dots, Y(S_n)] \quad (22.95)$$

$$Y(S_k) = T_e(S_k) \text{ or } I_1(S_k) \text{ or } P_1(S_k) \quad (22.96)$$

where  $S_1, \dots, S_n$  are the  $n$  values of slip considered for steady state. For transients, the test data  $I_{d1}(t)$ ,  $I_{q1}(t)$  are acquired at  $n$  time instants  $t_1, \dots, t_n$ . The stator current d-q components in synchronous coordinates now constitute the observation vector:

$$|Y| = [Y(t_1), Y(t_2), \dots, Y(t_n)] \quad (22.97)$$

**Table 22.4 Estimated parameters for various observation vectors [22]**

|                                                                           | Estimated parameters [ $\Omega$ ] |        |          |          |          |
|---------------------------------------------------------------------------|-----------------------------------|--------|----------|----------|----------|
|                                                                           | $R_2$                             | $R_3$  | $X_{l2}$ | $X_{l3}$ | $X_{l1}$ |
| Values used to calculate the input (theoretical) characteristics and runs | 3.1144                            | 0.2717 | 3.0187   | 4.6098   | 3.1316   |
| Values obtained after estimation from $T_e = f(S)$                        | 3.1225                            | 0.2717 | 3.032    | 4.6145   | 3.1261   |
| Values obtained after estimation from $I_s = f(S)$                        | 3.0221                            | 0.2722 | 2.868    | 4.5552   | 3.1946   |
| Values obtained after estimation from $i_a = f(S)$                        | 3.0842                            | 0.2718 | 2.9706   | 4.5931   | 3.1513   |
| Values obtained after estimation from $ i_s  = f(t)$                      | 3.0525                            | 0.2721 | 2.9099   | 4.5671   | 3.1762   |
| Values obtained after estimation from $T_e = f(t)$                        | 3.1114                            | 0.2664 | 3.0693   | 4.5831   | 3.093    |

$$Y(t_K) = I_{dl}(t_K) \text{ or } I_{ql}(t_K) \quad (22.98)$$

In general, the problem to solve may be written as

$$Y = f(\zeta_K, X) + \xi_K = 1, \dots, n \quad (22.99)$$

with  $\zeta = S_K$  for steady-state estimation and  $\zeta = t_K$  for transients estimation. The parameters  $\xi_K$  is the error for the  $K^{\text{th}}$  value of  $S_K$  (or  $t_K$ ).

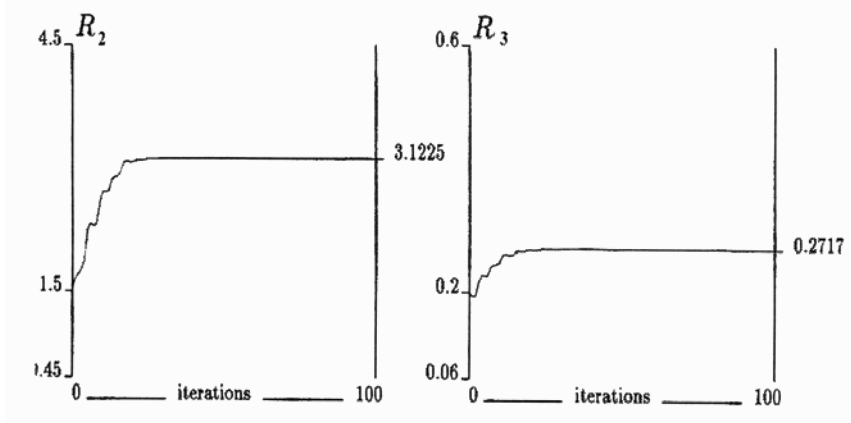


Figure 22.25 Convergence progress in rotor resistance estimation from  $T_e(S)$  [22]

There are many ways to solve such a problem iteratively. Gauss–Newton method, when used after linearization, tends to run into almost singularity matrix situations. On the contrary, the step-wise regression method, based on Gauss – Jordan pivot concept [21], seems to produce safely good results across the board. For details on the computation algorithm, see Reference [22]. The method's convergence should not depend on the width of change limits of parameters or on their assigned start values.

Typical results obtained in [22] are given in Table 22.4 for steady state and, respectively, for dynamic (direct starting) operation modes.

Sample results on convergence progress for  $T_e(S)$  observation vector show fast convergence rates in Figure 22.25. [22]

The corresponding torque/slip curve fitting is, as expected, very good (Figure 22.26).

Similarly good agreement for the torque–time curve during starting transients is shown in Figure 22.27 when the  $T_e(t)$  function is observed.

A few remarks are due.

- The curve fitting of the observed vectors over the entire explored zone is very good and the rate of convergence is safe and rather independent of the extension of parameter existence range.
- The estimated parameter vectors (Table 22.4) have only up to 3% variations depending on the observation vector; torque or current. This

is a strong indication that if the torque/slip is observed, the current/slip curve will also fit rather well, etc.

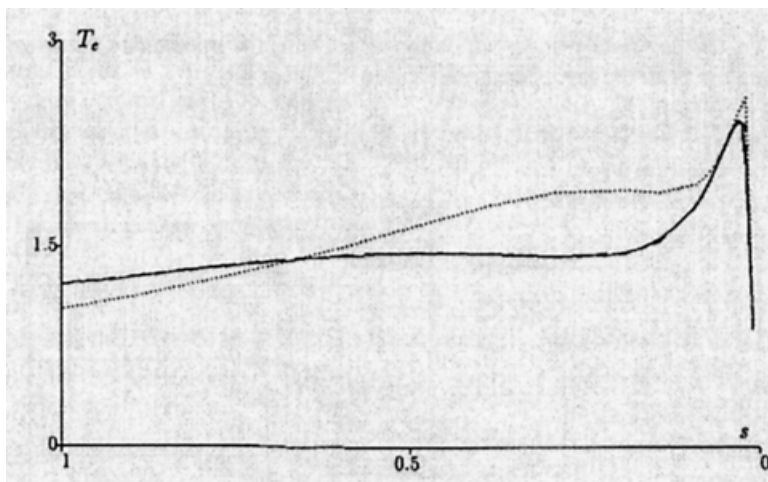


Figure 22.26 The static characteristic of electromagnetic torque  $T_e = f(S)$ : [22]

\_\_\_\_\_ – the theoretical characteristic of  $T_e = f(S)$   
 ..... – the characteristic calculated for starting values of parameters  
 ----- – the characteristic calculated for final values of estimated parameters

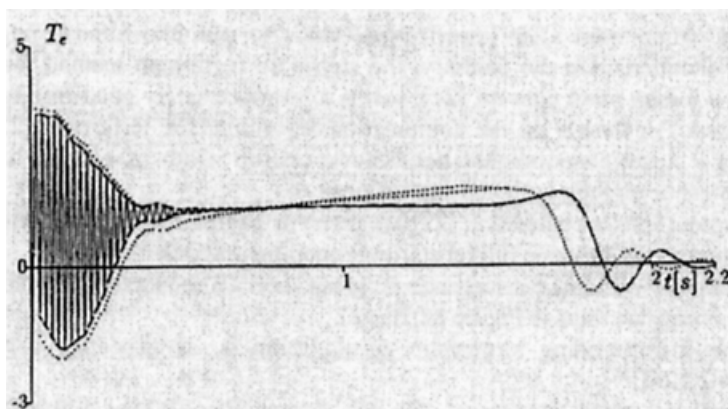


Figure 22.27 Runs of the electromagnetic torque  $T_e$  during the start – up: [22]

\_\_\_\_\_ – the theoretical run of  $T_e = f(t)$   
 ..... – the run calculated for starting values of parameters  
 ----- – the run calculated for final values of estimated parameters

- The phase current measurement is straightforward but the torque measurement is not so. We may however calculate the torque off line from measured voltages and currents  $V_a, V_b, I_a, I_b$ ,

$$T_e = \frac{3}{2} p_1 (\Psi_\alpha I_\beta - \Psi_\beta I_\alpha) \quad (22.100)$$

$$\begin{aligned} V_\alpha &= V_a \\ V_\beta &= \frac{1}{\sqrt{3}} (V_a + 2V_b) \\ I_\alpha &= I_a \\ I_\beta &= \frac{1}{\sqrt{3}} (I_a + 2I_b) \end{aligned} \quad (22.101)$$

$$\begin{aligned} V_\alpha &= \int (V_\alpha - R_1 I_\alpha) dt \\ V_\beta &= \int (V_\beta - R_1 I_\beta) dt \end{aligned}$$

The precision of torque estimates by (22.100) is good at rated frequency but smaller at low frequencies  $f_1$ . The integrator should be implemented as a filter to cancel the offset. In (22.100), the reduction of the electromagnetic torque due to core loss, which is smaller than 1% of rated torque in general, is neglected. If core loss  $p_{\text{core}}$  is known from the no load test, the torque of (22.100) may be reduced by  $\Delta T_{\text{ecore}}$ .

$$\Delta T_{\text{ecore}} = p_{\text{core}} \times \frac{p_1}{\omega_1} \quad (22.102)$$

- The direct start is easy to perform, as only two voltages and two currents are to be acquired and the data is processed off line. It follows that this test may be the best way to estimate the IM parameters for a large spectrum of operation modes. Such a way is especially practical for medium–large power IMs.
- Very large power machines direct starting is rather slow as the relative inertia is large and the power grid voltage drops notably during the process. In such a case, the method of parameter estimation data may be simplified as shown in the following paragraph where also the inertia is estimated.

#### 22.4.6 Large IM inertia and parameters from direct starting acceleration and deceleration data

Testing large power IMs takes time and notable human effort. Any attempt to reduce it while preserving the quality of the results should be considered.

Direct connection to the power grid, even if through a transformer at lower than rated voltage in a limited power manufacturer's laboratory, seems a practical way to do it. The essence of such a test consists of the fact that the acceleration would be slow enough to validate steady state operation mode approximation all throughout the process. In large power IMs, reducing the



supply voltage to 60% would do it in general. We assume that the core loss is proportional to voltage squared. By performing no load tests at decreasing voltages, we can separate first the mechanical from iron losses by the standard loss segregation method. The mechanical losses vary with speed. To begin, we may suppose a linear dependence

$$\begin{aligned} p_{\text{mec}} &\approx p_{\text{mec}0} \cdot \frac{n}{f_1/p} \\ p_{\text{iron}} &= p_{\text{iron}0} \cdot \left( \frac{V}{V_0} \right)^2 \end{aligned} \quad (22.103)$$

During the no load acceleration test, the voltage and current of two phases are acquired together with the speed measurement. Based on this, the average power is calculated. To simplify the process of active power calculation, synchronous coordinates may be used.

$$\begin{vmatrix} V_d \\ V_q \end{vmatrix} = \frac{2}{3} \begin{vmatrix} \cos(-\omega_1 t) & \cos\left(-\omega_1 t + \frac{2\pi}{3}\right) & \cos\left(-\omega_1 t - \frac{2\pi}{3}\right) \\ \sin(-\omega_1 t) & \sin\left(-\omega_1 t + \frac{2\pi}{3}\right) & \sin\left(-\omega_1 t - \frac{2\pi}{3}\right) \end{vmatrix} \cdot \begin{vmatrix} V_a \\ V_b \\ -(V_a + V_b) \end{vmatrix} \quad (22.104)$$

The same formula is valid for the currents  $I_d$  and  $I_q$ . Finally, the instantaneous active power  $P_1$  is

$$P_1 = \frac{3}{2} (V_d I_d + V_q I_q) \quad (22.105)$$

As  $V_d$ ,  $V_q$ ,  $I_d$ ,  $I_q$ , for constant frequency  $\omega_1$  are essentially d.c. variables the power signal  $P_1$  is rather clean and no averaging is required.

We might define now the kinetic useful power  $P_K$  as

$$\begin{aligned} P_K(n) &= P_{\text{elm}}(1-S) + p_{\text{mec}} \\ P_{\text{elm}} &= P_1 - 3R_s I_1^2 - p_{\text{iron}} \end{aligned} \quad (22.106)$$

As all terms in (22.106) are known for each value of speed, the kinetic useful power  $P_K(n)$ , dependent on speed, may be found. The time integral of  $P_K(n)$  gives the kinetic energy of the rotor  $E_c$ .

$$E_c = \int_0^t P_K(n) dt = \frac{J}{2} 4\pi^2 n_f^2 \quad (22.107)$$

where  $n_f$  is the final speed considered.

We may terminate the integration at any time (or speed) during acceleration. Alternatively,

$$P_K = J 4\pi^2 n \frac{dn}{dt} \quad (22.108)$$

At every moment during the acceleration process, we may calculate  $P_K$ , measure speed  $n$  and build a filter to determine  $dn/dt$ .

Consequently, the inertia  $J$  may be determined either from (22.107) or from (22.108) for many values of speed (time) and take an average.

However, (22.107) avoids the derivative of  $n$  and performs an integral and thus the results are much smoother.

Sample measurements results on a 7500 kW, 6000 V, 1490 rpm, 50 Hz, IM with  $R_1 = 0.0173\Omega$ ,  $P_{mec0} = 44.643$  kW and  $P_{iron} = 78.628$  kW are shown in Figure 22.28. [23]

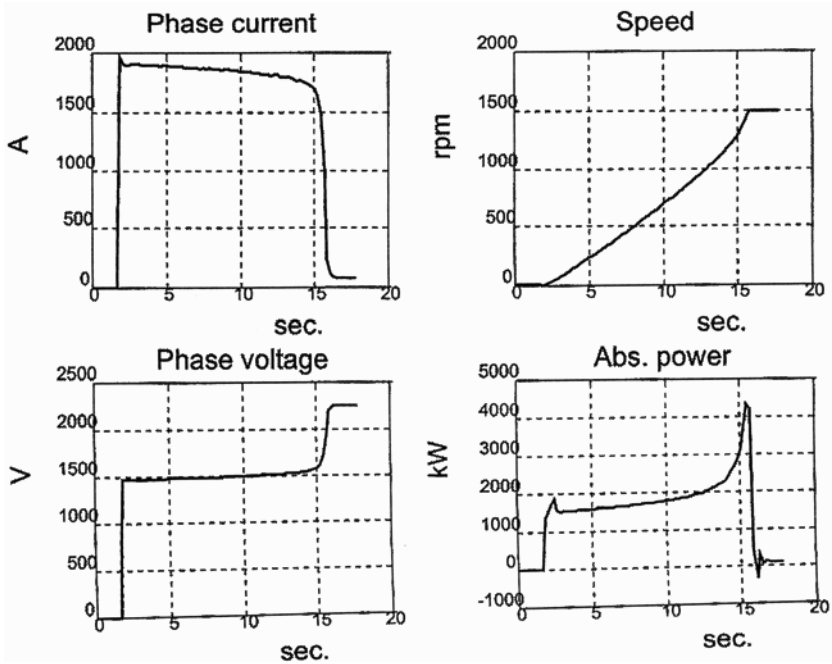


Figure 22.28 Current, phase, voltage, power and speed during direct starting of a 7500kW IM

The moment of inertia, calculated during the acceleration process, by (22.108), is shown in Figure 22.29.

Even with  $dn/dt$  calculated,  $J$  is rather smooth until we get close to the settling speed zone, which has to be eliminated when the average  $J$  is calculated.

The torque may also be calculated from two different expressions (Figure 22.30).

$$T = 2\pi J \frac{dn}{dt} \quad (22.109)$$

$$T_e = \frac{P_K}{2\pi n} \quad (22.110)$$

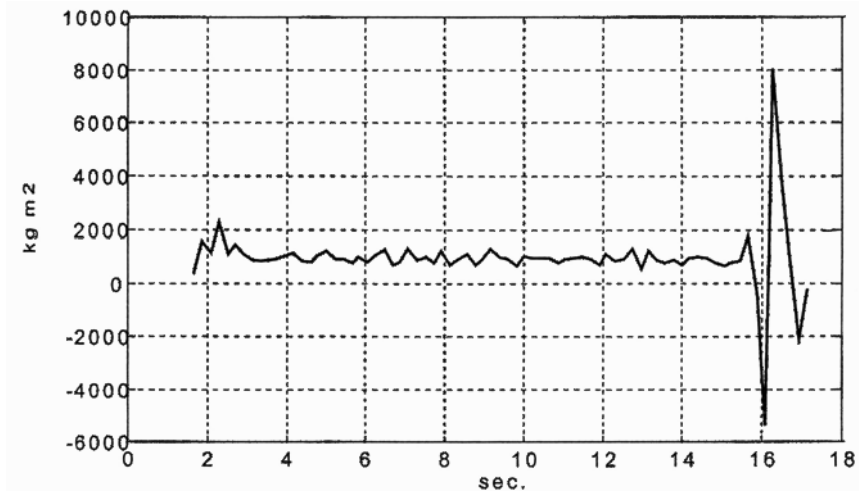


Figure 22.29 Inertia calculated during the acceleration process [23] using the speed derivative (22.108)

Again, Equation (22.110), which avoids  $dn/dt$ , produces a smoother torque/time curve.

The torque may be represented as a function of speed also, as speed has been acquired.

The torque in Figure 22.30 is scaled up to 6kV line voltage (star connection), assuming that the torque is proportional to voltage squared.

The testing may add a free deceleration test from which, with inertia known from the above procedure, the more exact mechanical loss dependence on speed  $p_{mec}(n)$  may be determined.

$$P_{mec}(n) = -J2\pi \frac{dn}{dt} \quad (22.111)$$

Then  $P_{mec}(n)$  from (22.111) may be used in the acceleration tests processing to get a better inertia value. The process proved to converge quickly. Finally, an exponential regression of mechanical losses has been adopted.

$$P_{mec} = \alpha n^{2.4}; \alpha = 0.001533 \quad (22.112)$$

The stator voltage attenuation during free deceleration has also been acquired.

Sample deceleration test results are shown in Figure 22.31 a,b,c.

The mechanical losses, determined from the deceleration tests, have been consistently smaller (that is for quite a few machines tested: 800 kW, 2200 kW, 2800 kW, 7500 kW) than those obtained from loss segregation method.

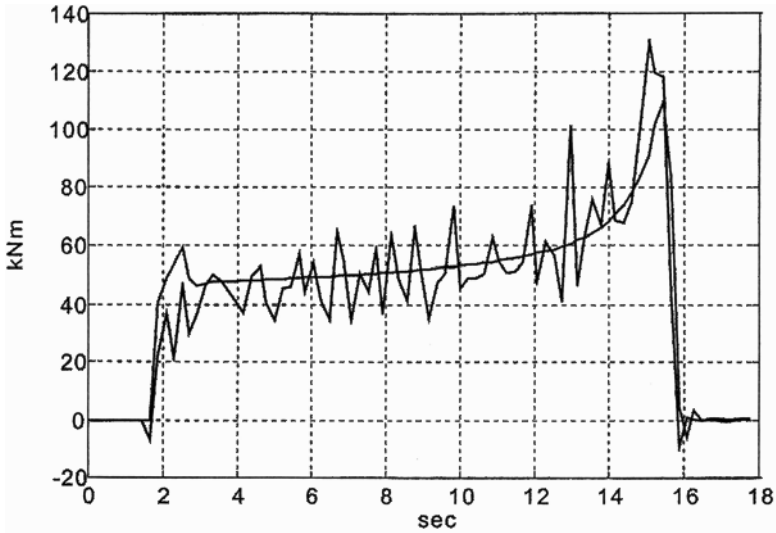


Figure 22.30 Torque versus time during acceleration [23] from (22.109) and from (22.110)

In our case at no load speed  $P_{mec} = 64.631 \text{ kW}$  from the free deceleration tests. That is, 14 kW less than from the loss segregation method.

This difference may be attributed to additional (stray losses).

To calculate the stray load losses, we may adopt the formula

$$P_{\text{stray}}(I) = \left[ (P_{\text{mec}})_{\text{l.s.m.}} - (P_{\text{mec}})_{\text{f.d.m.}} \right] \cdot \left( \frac{I}{I_0} \right)^2 \quad (22.113)$$

l.s.m. – loss segregation method

f.d.m. – free deceleration method

where  $I$  is the stator current and  $I_0$  the no load current.

To complete loss segregation and secure performance assessment, all IM parameters have to be determined. We may accomplish this goal avoiding shortcircuit test due to both time (work) needed and also due to the fact that large IMs tend to have strong rotor skin effect at standstill and thus a double cage rotor circuit has to be adopted.

The IM parameters may be identified from the acceleration and free deceleration test data alone through a regression method. [23]

## 22.5. NOISE AND VIBRATION MEASUREMENTS: FROM NO-LOAD TO LOAD

Noise in induction machine has electromagnetic and mechanical origins. The noise level accepted depends on the environment in which the IM works. At

the present time, standards (ISO1680) prescribe noise limits for IMs, based on sound power levels measured at no load conditions, if the noise level does not vary with load. It is not as clear how to set the sufficient conditions for no load noise tests.

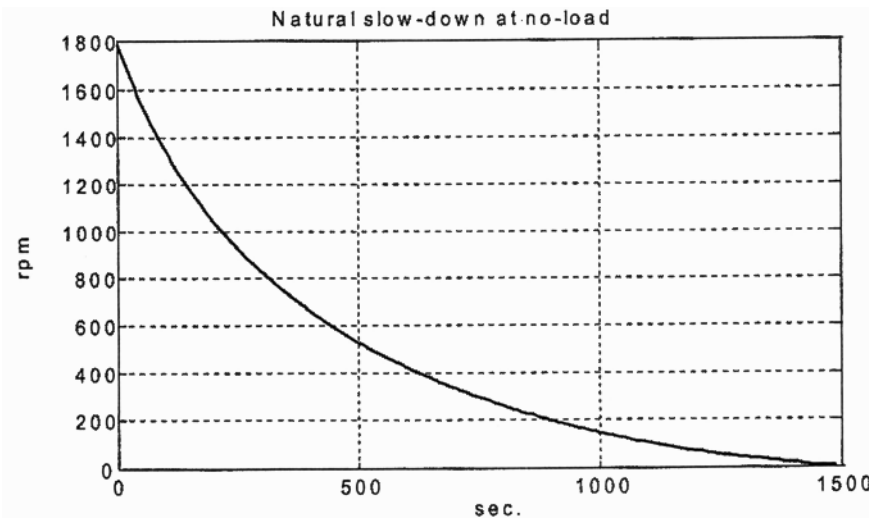
The load machine produces additional noise and thus, it is simpler to retort to IM no load noise tests only.

If the electromagnetic noise is relatively large, it is very likely that under load it will increase further. [22] Also, when PWM converter fed, the IM noise will change from no-load to load conditions. Skipping the basics of sound theory [26] in electrical machines, we will synthesize here ways to measure the noise under load in IMs.

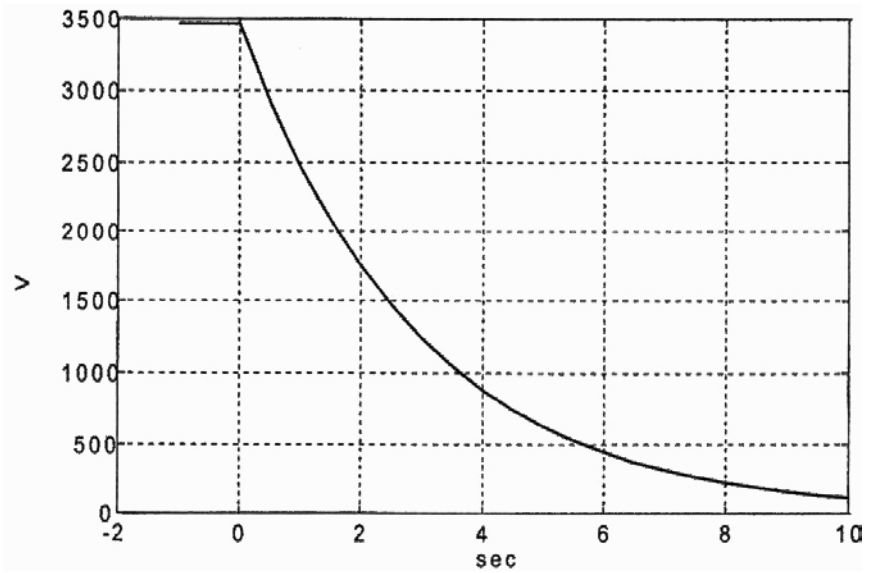
**22.5.1 When on-load noise tests are necessary?**

Noise tests with the machine on no-load (free at shaft) are much easier to perform, eventually in an acoustic chamber (anechoic room for free field or reverberant room for diffuse field). So it is natural to take the pains and perform noise measurements on load only when the difference in noise level between the two situations is likely to be notable.

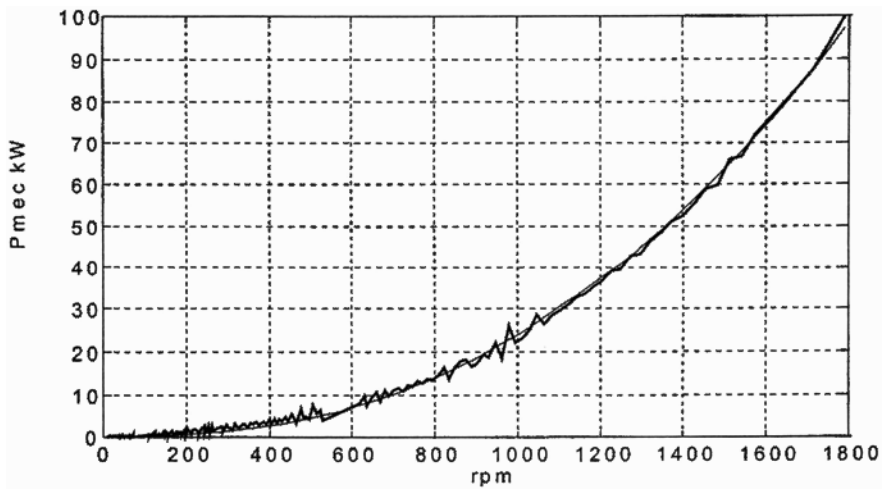
An answer to this question may be that the on load noise measurements are required when the electromagnetic noise at no-load is relatively important. Under load, space harmonics fields accentuate and are likely to produce increased noise, aside from the noise related to torque pulsations (parasitic torques) and uncompensated radial forces.



a.)



b.)



c.)

Figure 22.31 Free deceleration tests [23]

a.) speed versus time; b.) stator voltage amplitude versus time  
c.) mechanical losses versus speed

To separate the no-load electromagnetic noise sound power level  $L_{W,em}$ , three tests on no load are required [27,29]:

a) The no-load sound power  $L_{W0}$  is measured according to standards.

- b) After taking off the ventilator to sound power level  $L_{Wf}$  is measured. Subtracting  $L_{Wf}$  from  $L_{Wo}$ , we determine the ventilator system sound power level  $L_{W,vent}$ .

$$L_{W,vent} = 10 \log(10^{0.1L_{Wo}} - 10^{0.1L_{Wf}}) \quad (22.114)$$

- c) Without the ventilator, the no-load noise measurements are carried out at ever smaller voltage level before the speed decreases. By extrapolating for zero voltage the mechanical sound power level  $L_{W,mec}$  is obtained. Consequently, the electromagnetic sound level at no load  $L_{W,em}$  is

$$L_{W,em} = 10 \log(10^{0.1L_{Wo}} - 10^{0.1L_{W,mec}}) \quad (22.115)$$

Now if the electromagnetic sound level  $L_{W,em}$  is more than 8dB smaller than the total mechanical sound power, the noise measurements on load are not necessary.

$$L_{W,em} < 8\text{dB} + 10 \log(10^{0.1L_{W,mec}} + 10^{0.1L_f}) \quad (22.116)$$

## 22.5.2 How to measure the noise on-load

There are a few conditions to meet for noise tests

- The sound pressure has to be measured by microphones placed in the acoustic far field (about 1m from the motor)
- The background sound pressure level radiated by any sound source has to be 10 dB lower than the sound pressure generated by the motor itself
- Every structure-based vibration has to be eliminated

Acoustical chambers fulfil these conditions but large machine shops, especially in the after hours, are also suitable for noise measurements.

Under on-load conditions, the loading machine and the mechanical transmission (clutch) produce additional noise. As the latter is located close to the IM, its contribution to the resultant noise power level is not easy to segregate.

Separation of the loading machine may be performed through putting it outside the acoustic chamber which contains the tested motor.

Also, an acoustic capsule may be built around the coupling plus the load machine with the tested motor placed in the machine shop.

Both solutions are costly and hardly practical unless a special “silent” motor is to be tested thoroughly.

Thus, in situ noise measurements, in large machine shops, after hours, appears as the practical solution for noise measurements on load.

The sound pressure or power is measured in the far field 1 m from the motor in general. However, to reduce the influence of the loading machine noise, the measurements are taken at near field 0.2 m from the motor.

However, in this case, near field errors occur mainly due to motor vibration modes.

To eliminate the near field errors, comparative measurements in the same points are made.

For example, in a near field point (0.2m from the motor), the sound pressure is measured on no-load and with the load coupled to obtain  $L_{W,on}$  and  $L_{W,n}$ . The difference between the two represents the load contribution to noise and is not affected by the near field error.

$$\Delta L_n = L_{W,n} - L_{W,on} \quad (22.117)$$

This method of near field measurements is most adequate for tests at the working place of the IMs.

Noise, in the far or near field, may be measured directly by sound pressure microphones (Figure 22.32) or calculated based on vibration speed measurements.

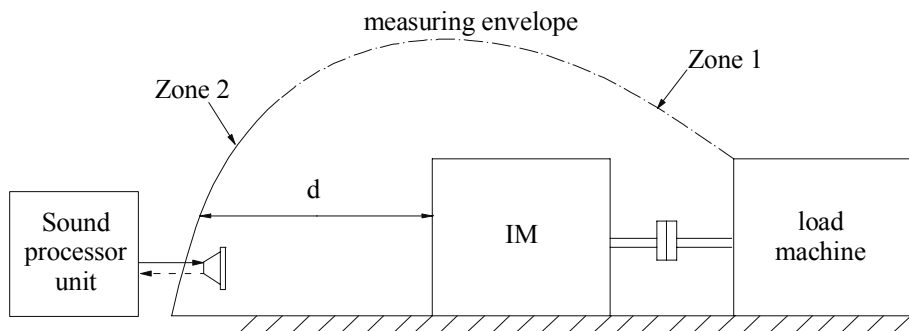


Figure 22.32 Sound intensity measurement arrangements

The new microphone method measures sound pressure and produces very good results if the microphone placement avoids the aggregation of load machine noise contributions. Their placements in zone 2 (Figure 22.32) at  $d = 1$  m from motor, though the best available solution may still lead to about 3 dB increase in noise as loading machine contribution. This may be considered a systematic error at worst.

The artificial loading by PWM dual frequency methods avoids the presence of the load machine and it may constitute a viable method to verify IM noise measurements under load.

The noise level may also be calculated from vibration mode speeds measurement. Measurements of IM frame vibration mode speeds are made on its frame. This is the traditional method and is described in detail in the literature. [24–28]

Experimental results reported in Ref [27] lead to conclusions such as:



- The sound power level may be determined directly by sound intensity (pressure) measurement in the far field (1 m away from IM) through microphones (Figure 22.32)
- The sound power level of the IM under load may be determined indirectly via measurements by microphones, on no-load and on-load, made in the near field (0.2 m) of the IM (Equations (22.125)–(22.126))
- Finally, the sound power may be calculated from noise or vibration measurements depending on which the mechanical or electromagnetic noise, dominates
- All these procedures seem worthy of further consideration, as their results are close. In [27], the far field microphone method departed from the other two by 0.80 dB (from 78.32 dB to 79.12 dB on-load)
- A  $\Delta L_n$  of 6.84dB increase due to load was measured for a 5kW IM in Ref [27]

For details on noise and vibration essentials, see [28].

In the power electronics era, IM noise theory testing and reduction are expected to prompt new R & D efforts.

## 22.6 SUMMARY

- IM testing refers essentially to loss segregation, direct and indirect load testing, parameter estimation and noise and vibration measurements.
- Loss segregation is performed to verify design calculations and avoid full load testing for efficiency calculation.
- The no-load and standstill tests are standard in segregating the mechanical losses, fundamental core losses, and winding losses.
- The key point in loss segregation is however the so-called stray load losses.
- Stray load losses may be defined by identifying their origins: rotor and stator surface and teeth flux pulsation space harmonics core losses, the rotor cage non-fundamental losses. The latter are caused by stator m.m.f. space harmonics, augmented by airgap permeance harmonics, and/or inter-bar rotor core losses. Separating these components in measurements is not practical.
- In face of such a complex problem, the stray load losses have been defined as the difference between the total losses calculated from load tests and the ones obtained from no-load and standstill (loss segregation) tests.
- Though many tests to directly measure the stray load losses have been proposed, so far only two seem to hold.
- In reverse motion test, the IM fed at low voltage is driven at rated speed but in reverse direction to the stator voltage system. The slip  $S \approx 2$ . Though the loss equivalence with the rated load operation is far from complete, the elimination of rotor cage fundamental loss from calculations “saves” the method for the time being.

- In the no-load test extension over rated voltage – so far performed only on small IMs – the power input minus stator winding losses depart from its proportionality to voltage squared. This difference is attributed to additional “stray load” losses considered then proportional to current squared.
- The stator resistance is traditionally d.c. measured. However, for large IMs, a.c. measurements would be useful. The a.c. test at low voltage with the rotor outside the machine is considered to be pertinent for determining both the stator resistance and leakage inductance.
- The no-load and standstill tests may be performed by using PWM converter supplies, especially for IM destined for variable speed drives. Additional losses, in comparison with sinusoidal voltage supply, occur at no-load, not so at standstill test as the switching frequency in the converter is a few kHz.
- If the distribution of the losses is known, in the moment their source is eliminated by turning off the stator winding, the temperature gradient is proportional to those losses. Consequently, measuring temperature versus time leads to loss value. Unfortunately, loss distribution is not easy to find unless FEM is used and the temperature sensors, placed at sensitive points, represent an intrusion which may be accepted only for prototyping.
- On the contrary, the calorimetric method – in the dual chamber version – leads to the direct determination of overall losses, with an error of less than 5% generally.
- Load tests are performed to determine efficiency and temperature rise for various loads.
- Efficiency may be calculated based on loss segregation methods or by direct (or indirect: artificial) loading tests. In the latter case, both the input and output powers are measured.
- Standard methods to determine efficiency are amply presented in the international standards IEEE-112 (USA), IEC-34-2 (Europe) and JEC (Japan). The essential difference between these three standards lies in the treatment of non-fundamental (or stray load) losses. JEC apparently considers the stray load losses as zero, IEC-34-2 takes them as a constant: 0.5% of rated power. IEEE-112, through its many alternative methods, determines them as the difference between overall losses in direct load tests and no-load losses at rated voltage plus standstill losses at rated current. Alternatively, the latter sets constant values for stray load losses, all higher than 0.5% - from 1.8% below 90 kW to 0.9% above 1800 kW. So, in fact, the same IM tested according to these representative standards would have different labelled efficiency. Or, conversely, an IM with the same label efficiency and power will produce most power if the efficiency was measured according to IEEE-112 and least power for JEC efficiency definition. The differences may run as high as 2%.

- IEEE-112B by its provisions seems much closer to reality.
- The direct torque measurement through a brushless torque-meter in direct load tests may be avoided either by a calibrated d.c. generator loading machine or by estimating the torque of an inverter-fed IM machine load. Alternatively, the tested IM may be run both as a motor and generator at equal slip values. Rather, simple calculations will allow stray load loss and efficiency computation based on input power, current, voltage, and slip precise measurements.
- Avoiding altogether the burden of coupling a second machine at the tested IM shaft led to artificial (indirect) loading. The two frequency method introduced in 1921 has been used only for temperature rise calculations as loss equivalence with direct load tests was not done equitably.
- The use of PWM converter to mix frequency and provide indirect (artificial) loading apparently changes the picture. Equivalence of losses may be now pursued elegantly and the whole process may be mechanized through PC control. Good agreement in efficiency with the direct methods has been reported recently. Is this the way of the future? More experience with this method will only tell.
- A rather conventional indirect method, applied by a leading manufacturer for 40 years is the forward shortcircuit approach. In essence, the unsupplied IM is first driven at rated speed by a 10% rating IM of the same number of poles. Then another 10% rating motor drives a synchronous generator to produce about 80–85% rated frequency of the tested IM. Finally, the a.c. generator is excited and its terminals are connected to the tested motor. The voltage is raised to reach the desired current. The tested IM works at high slip as a generator and at rated speed and thus the active power output is low: 10% of rated power. Despite the fact that the distribution of losses differs from that in direct load tests, with more losses on rotor and less on stator and skin rotor effect presence, good stator temperature rise agreement with direct load tests has been reported for large IMs. The introduction of bi-directional power electronics to replace the second drive motor plus a.c. generator might prove a valuable way to improve on this genuine and rather practical method for large power IMs.
- Parameter estimation is essential for IM precise modelling over a wide spectrum of operation modes.
- Single and multiple rotor loop circuit models are used for modelling.
- Basically, only no-load and shortcircuit tests are standardized for parameter estimation.
- No-load tests provide reliable results on the core loss resistance and magnetization inductance dependence on magnetization current.
- Shortcircuit tests performed at standstill and at rated frequency contain too much skin effect influence on rotor resistance and leakage

inductance, which precludes their usage for calculating on-load IM performance.

- Even for starting torque and current estimation at rated voltage, they are not proper as they are performed up to rated current and thus the leakage flux path saturation is not considered. Large errors at full voltage – up to 60–70% – may be expected in torque in extreme cases.
- Two frequency shortcircuit tests at rated current – one at rated frequency and the other at 5% rated frequency – could lead to a dual cage rotor model good both for large and low slip values, provided the current is not larger than 150–200% of rated current; most PWM converter-fed general variable speed IM drives qualify for these conditions.
- The standstill frequency response (SSFR) method is an improvement on the two-frequency method but, being performed at low current and many frequencies, excludes leakage saturation effect as well and time prohibitive. Its validity is restricted to small deviation transients and stability analysis in large IMs.
- Even catalogue – data – based methods have been proposed to match (by a single model with dual virtual cage rotor) both the starting and rated torque and currents. However, in between, around breakdown torque, large errors persist.
- The general regression method, based on steady state or transients measurements, for wide speed range, from standstill to no-load speed, has been successfully used to estimate the parameters of an equivalent dual cage rotor model according to the criterion of least squared error. Thus, a wide spectrum of operation modes can be handled by a single but adequate model. A slow acceleration start and a direct (fast) one are required to cover large ranges of current and rotor slip frequency. The slow start may be obtained by mounting an inertia disk on the shaft. However direct fast start transients tend to produce, by the general regression method, about the same parameters as the steady state test and may be preferred.
- A simplified version of the naturally slow acceleration (quasi-steady state electromagnetic-wise) for very large IMs has been successfully used to separate the losses, calculate the inertia and dual cage rotor model parameters that fit the whole slip range.
- Noise and vibration tests are environmental constraints. Only no-load noise tests are standardized but sometimes on-load noise is larger by 6–8 dB or more
- When and how on-load in situ noise tests are required (done) is discussed in detail. For PWM converter fed IMs, variable speed noise tests seem necessary. [29]
- We did not exhaust, rather prioritize, the subject of IM testing. Many other methods have been proposed and will continue to be introduced

as the IM experimental investigations may see a new boost by making full use of PWM converter variable voltage and frequency supplies.

- An interesting example is the revival of the method of driving the IM on no-load through the synchronous speed and detect the power input jump at that speed. This power jump corresponds basically to hysteresis losses. The test is made easier to do with a PWM converter supply with a precise speed control loop.

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